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Permanent-Magnet-Assisted Reluctance Synchronous Starter/Alternators for Electric Hybrid Vehicles

8.1	Introduction	8-1
8.2	Topologies of PM-RSM	8-3
8.3	Finite Element Analysis.....	8-5
	Flux Distribution • The $d-q$ Inductances • The Cogging Torque • Core Losses Computation by FEM	
8.4	The $d-q$ Model of PM-RSM	8-12
8.5	Steady-State Operation at No Load and Symmetric Short-Circuit.....	8-19
	Generator No-Load • Symmetrical Short-Circuit	
8.6	Design Aspects for Wide Speed Range Constant Power Operation	8-21
8.7	Power Electronics for PM-RSM for Automotive Applications	8-27
8.8	Control of PM-RSM for EHV.....	8-30
8.9	State Observers without Signal Injection for Motion Sensorless Control.....	8-32
8.10	Signal Injection Rotor Position Observers.....	8-34
8.11	Initial and Low Speed Rotor Position Tracking.....	8-34
8.12	Summary.....	8-39
	References.....	8-41

8.1 Introduction

The permanent-magnet-assisted reluctance synchronous machine (PM-RSM), also called the interior permanent magnet (IPM) synchronous machine, with high magnetic saliency was proven to be competitive, price-wise (Figure 8.1) [1], and superior in terms of total losses (Figure 8.2) [2] for starter/alternator automobile applications where a large constant power speed range $\omega_{max}/\omega_b > 3$ to 4 is required.

The cost comparisons show the PM-RSM starter alternator system, including power electronics control, to be notably less expensive than the surface PM synchronous machine or switched reluctance machine system at the same output. It is comparable with the cost of the induction machine system. In terms of

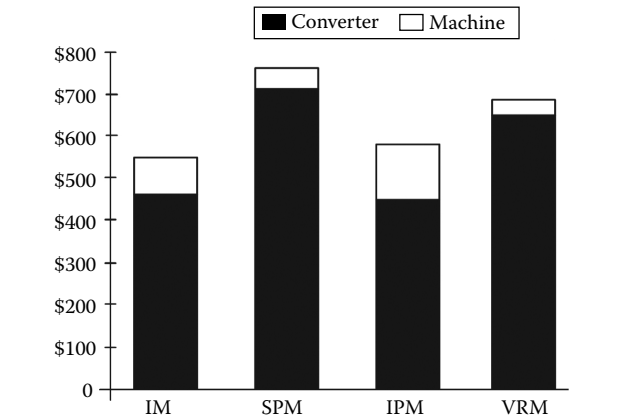


FIGURE 8.1 Cost comparisons for four alternators at 6 kW and 42 Vdc.

total machine plus power electronics losses, the PM-RSM is slightly superior even to the surface PM synchronous machine (Figure 8.2), and notably superior to the induction machine system, all designed for the same machine volume, at 30 kW [2].

It was also demonstrated that PM-RSM [2] is capable of a notably larger constant power speed range than the surface PM synchronous, or the induction, or the switched reluctance machine of the same volume.

In essence, both the lower cost and the wide constant power speed range are explained by the combined action of PMs and the high magnetic saliency torque to reduce the peak current for peak torque at low speed and reduce flux/torque at high speeds.

Starter/alternators for automobile applications are forced to operate at a constant power speed range $\omega_{max}/\omega_b > 3$ to 4 and up to 12 to 1. The higher the interval (without notably oversizing the machine or the converter) constant power speed range, the better.

This is how the PM-RSM becomes a tough competitor for electric hybrid vehicles (EHVs). The larger ω_{max}/ω_b is, the smaller the PM contribution to torque.

In what follows, we will treat the main topological aspects, field distribution, and parameters by finite element method (FEM), lumped parameter modeling of saturated PM-RSM, core loss models, design issues for wide ω_{max}/ω_b ratios, and system models for dynamics and vector flux-oriented control (FOC) and direct torque and flux control (DTFC) with and without position control feedback.

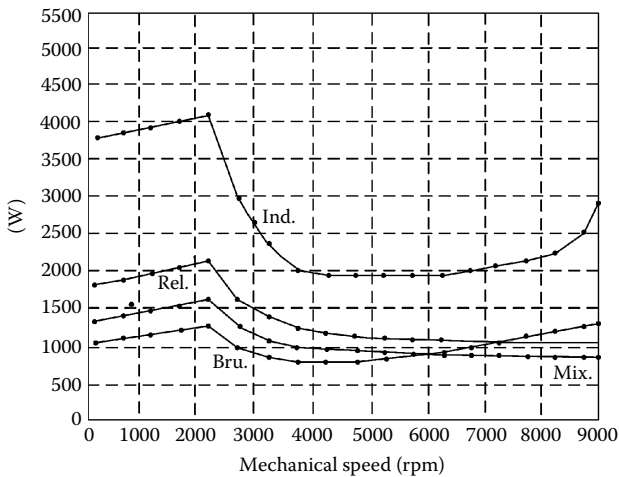


FIGURE 8.2 Loss comparisons between four starter motor-generator systems at 30 kW.

8.2 Topologies of PM-RSM

The PM-RSM is, in fact, a multilayer flux barrier and PM rotor synchronous machine — an interior multilayer PM rotor machine, in other words. The standard IPM (interior PM machine) has only one PM rotor layer per pole, and the PM contribution to torque is predominant (Figure 8.3a). In PM-RSMs, the high magnetic saliency created by the multiple flux barriers in the rotor make reactance torque predominant at low speeds when highest torque is required.

The stator core of the PM-RSM is provided with uniform slots that host a distributed ($q > 2$) three-phase winding with chorded coils. The rotor core may be built of conventional (transverse) laminations with stamped multiple flux barriers per pole, filled with PM layers (Figure 8.3b), or from axial laminations with multiple PM layers per pole (Figure 8.3c).

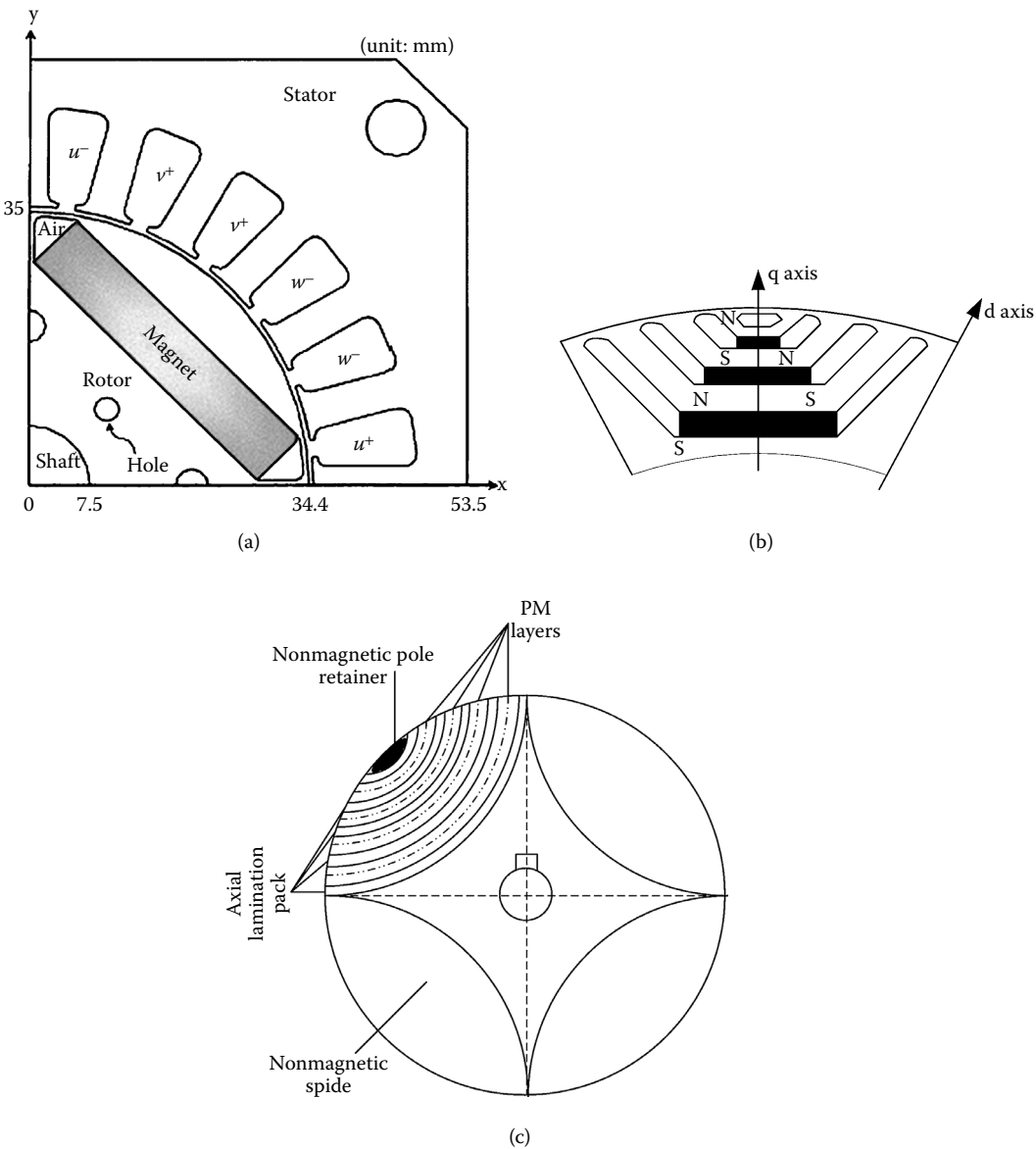


FIGURE 8.3 Rotor for permanent-magnet-assisted reluctance synchronous machine (PM-RSM): (a) standard interior permanent magnet (IPM), (b) with transverse laminations, and (c) with axial laminations.

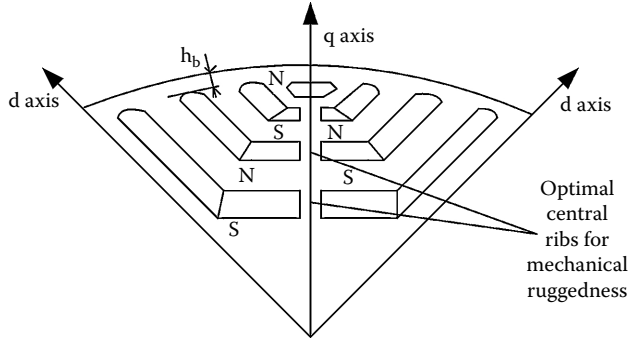


FIGURE 8.4 Mechanically acceptable flux barrier geometry.

The transverse lamination rotor with multiple flux barriers requires magnetic bridges to leave the lamination in one piece and provide enough mechanical resilience up to maximum design speed.

It turns out that mechanical speed limitations, due to centrifugal forces, basically, lead to magnetic bridges that are shorter (tangentially), while their thickness varies radially from 0.6 to 1 (1.2) mm minimum (Figure 8.4) [3]. The shorter flux bridges tend to result in larger magnetic permeance along the q axis (which is not desirable) but to larger permeance along the d axis also (which is desirable), and the L_{dm} to L_{qm} difference may be maintained acceptably large. L_{dm}/L_{qm} ratios up to 10/1 may be obtained this way (L_{dm} and L_{qm} are the magnetization inductances). Placing the PMs on the bottom of the flux barriers (Figure 8.3b) has the advantage that the former are more immune to demagnetization at peak current (or torque). However, it has the disadvantage that the PM flux leakage is larger, so more PM weight is required. Moreover, when the PMs are placed on the lateral part of the flux barriers the reverse is true. Depending on the overload specifications and constant power speed range, the PMs may be placed either on the bottom or on the lateral parts of the flux barriers.

The axially laminated rotor (Figure 8.3c) [4, 5] is made of overlaid layers of axial laminations and PM flexible ribbons for each pole. A nonmagnetic rotor spider sustains the poles, and a nonmagnetic pole retains them to the spider with nonmagnetic bolts.

There are no flux bridges between the axial lamination layers; thus, L_{qm} is notably smaller. Also, the PM flux loss (leakage) tangentially in the rotor is smaller. There are, however, three problems associated with the axially laminated anisotropic (ALA) rotor:

- The manufacturing is not standard.
- The mechanical rigidity is not high, so the maximum peripheral speed is limited to, perhaps, 30 m/sec.
- The flux harmonics due to the open flux barriers produce flux pulsations in the rotor, and the q axis armature reaction space harmonics create magnetic flux perpendicular to axial lamination plane. Both of these phenomena produce additional losses in the rotor, as the airgap is necessarily low to ensure high magnetic saliency: L_{dm}/L_{qm} .

Two of three radial slits in the rotor core (laminations) and the making of the PM of three (four) pieces axially will reduce those losses to reasonable values [4]. Saliencies $L_{dm}/L_{qm} = 25$ under saturated conditions were obtained with a two-pole 1.5 kW machine [4].

The magnetic saliency ratio L_{dm}/L_{qm} refers here only to the main flux path. The leakage inductance L_{sl} , when included, makes the ratio L_d/L_q smaller:

$$\frac{L_d}{L_q} = \frac{L_{dm} + L_{sl}}{L_{qm} + L_{sl}} < \frac{L_{dm}}{L_{qm}} \quad (8.1)$$

Increasing the number of PM layers/pole leads to increasing the L_{dm}/L_{qm} and $L_{dm} - L_{qm}$ but, above four PM layers (flux barriers) per half-pole, the improvements are minimal. Two to four flux barriers, depending on the rotor diameter and on the number of poles, seem to suffice [6].

Modification of the geometry of flux barriers, mainly their relative thickness, the shape of the flux bridges, together with the number of layers (two, three, or four), are tools to improve the L_{dm}/L_{qm} ratio and the $1 - \frac{L_{qm}}{L_{dm}}$ saliency index, and thus improve performance [7]. The higher the L_{dm}/L_{qm} , the larger the power factor and the constant power speed range, as the torque is proportional to the $1 - \frac{L_{qm}}{L_{dm}}$ factor.

On the other hand, the lower the L_{qm} , the higher the short-circuit current. So, high magnetic saliency (low L_{qm}/L_{dm}) means large short-circuit current levels.

8.3 Finite Element Analysis

The complex topology of the rotor, combined with the small airgap, stator slot openings, and the distributed stator windings, make a finite element analysis of flux distribution in the machine a practical way to grasp some essential knowledge about the PM-RSM capability in terms of torque for given geometry and stator magnetomotive force (mmf).

8.3.1 Flux Distribution

Consider the geometry given in Table 8.1 [8]. The stator-rotor structure is shown in Figure 8.5a, where the flux lines for zero stator currents are shown. As expected, the flux lines “spring” perpendicular from the PMs and then spread between the flux barriers radially in the rotor [8] and then through the stator teeth.

The PM airgap flux density distribution in Figure 8.5b shows the presence of both multiple PM layers and the stator slot openings. The average value is about 0.3 T, while the distribution is similar to that of no-load airgap flux density in induction machines.

Then, current of given amplitude $I\sqrt{2}$ is injected in phase A with $(-I/\sqrt{2})$ in phases B and C. This way, the stator mmf distribution axis falls along the phase A axis. The rotor axis d (of highest permeance) is then moved from the mmf axis by increasing mechanical angle.

Admitting a certain current density and slot filling factor, the slot mmf may be found without knowing the number of turns per coil in the two-layer, eight-poles, three-phase winding.

The calculation of torque, through FEM, is done for various maximum slot mmfs and angles, until the maximum torque reaches the target of 140 Nm as a certain power angle $\delta_{id} = \tan^{-1}(\frac{I_q}{I_d})$ (Figure 8.6a and Figure 8.6b).

The flux lines for the peak torque conditions, with $\delta_{id} = 46^\circ$ (electrical degrees), are shown in Figure 8.7. The airgap flux density for the same situation, shown in Figure 8.8, evidentiates the strong influence of the stator mmf contribution (see Figure 8.4) on airgap flux density. The question arises as to if, in such conditions, the PMs are not demagnetized. The flux density “getting out” from the lowest barrier

TABLE 8.1 Geometry of a 140 nm Peak Torque Permanent-Magnet-Assisted Reluctance Synchronous Machine (PM-RSM)

Parameter	Value
Stator outer diameter	245.2 mm
Rotor outer diameter	174.2 mm
Rotor inner diameter	113 mm
Airgap	0.4 mm
Stack length	68 mm
PM parameter at 20°C	$B_r = 0.87$ T $H = 0.66$ kA/m

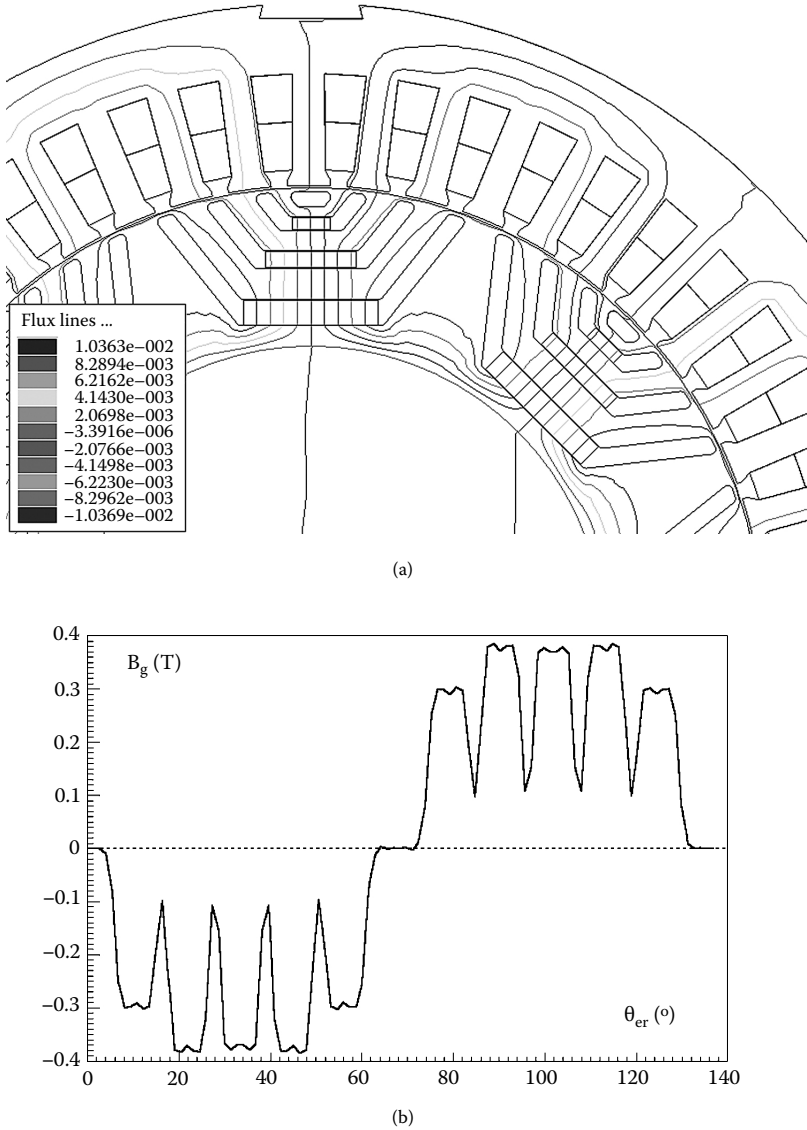


FIGURE 8.5 (a) Permanent magnet (PM) flux lines (zero stator current) and (b) airgap flux density distribution.

PM (Figure 8.3b) indicates clearly that the minimum flux density is 0.42 T and, thus, the PMs are safe (Figure 8.9).

The computation of flux density distribution in the stator teeth, in the middle or at their bottom [8], shows that the magnetic saturation may reach 1.85 T in the teeth and 1.65 T in the back iron, in this particular design. The resultant airgap flux for the maximum torque depicted in Figure 8.10 is not far from a sinusoid waveform.

8.3.2 The d - q Inductances

The above analysis indicates that even for rather high electric and magnetic loadings, the airgap flux distribution still keeps close to a sinusoid. Consequently, the orthogonal axis (d - q) circuit model still leads, at least for preliminary design or control purposes, to practical results. It goes without saying that varying the magnetic saturation has to be considered in the d - q model. Also, through FEM, we may

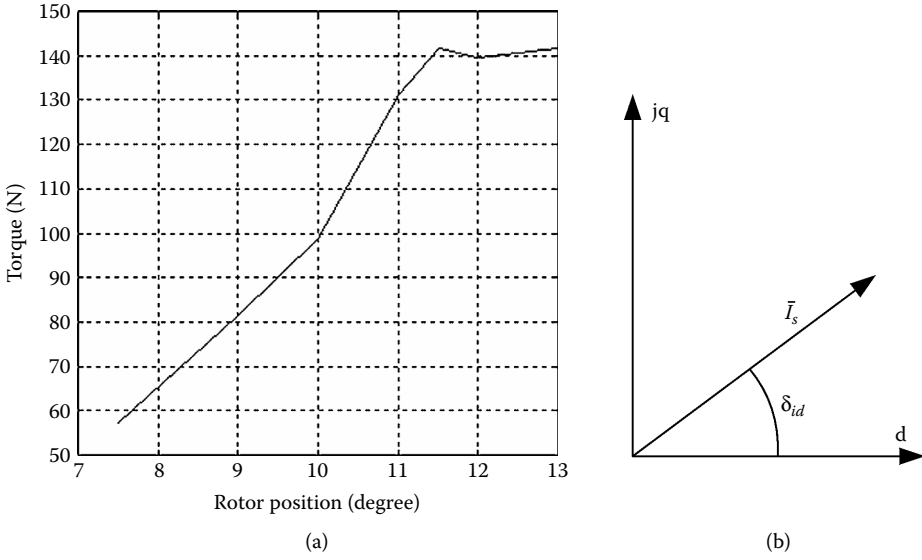


FIGURE 8.6 (a) The finite element method (FEM) calculated torque vs. rotor position for $n_c I\sqrt{2} = 500$ (Aturns/coil) in phase A, and $-n_c I/\sqrt{2}$ in phases B and C ($j_{\text{peak}} = 17.66$ A/mm²) and (b) power angle δ_{id} .

proceed and calculate, for given stator current, d - q current load angle, δ_{id} : the flux linkage in phases A, B, C: Ψ_A, Ψ_B, Ψ_C .

Based on these values, the d - q model flux linkages may be calculated:

$$\Psi_d + j\Psi_q = \frac{2}{3} \left(\Psi_A e^{-j\delta_{id}} + \Psi_B e^{-j\left(\delta_{id} - \frac{2\pi}{3}\right)} + \Psi_C e^{-j\left(\delta_{id} + \frac{2\pi}{3}\right)} \right) \quad (8.2)$$

$$i_d = I\sqrt{2} \cos \delta_{id} \quad i_q = -I\sqrt{2} \sin \delta_{id}$$

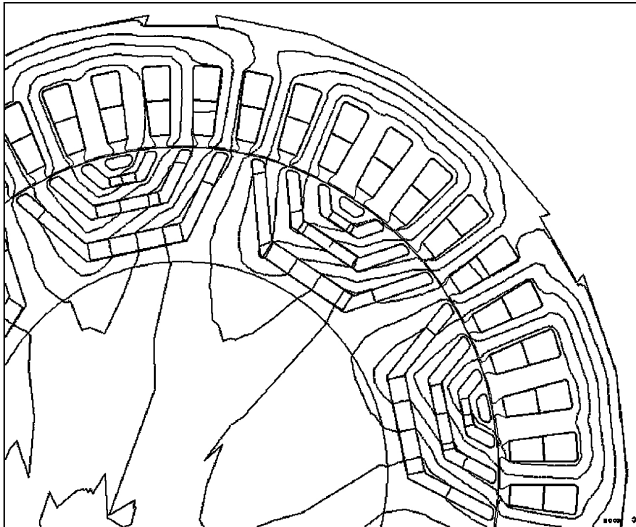


FIGURE 8.7 The flux lines for the maximum torque (140 nm).

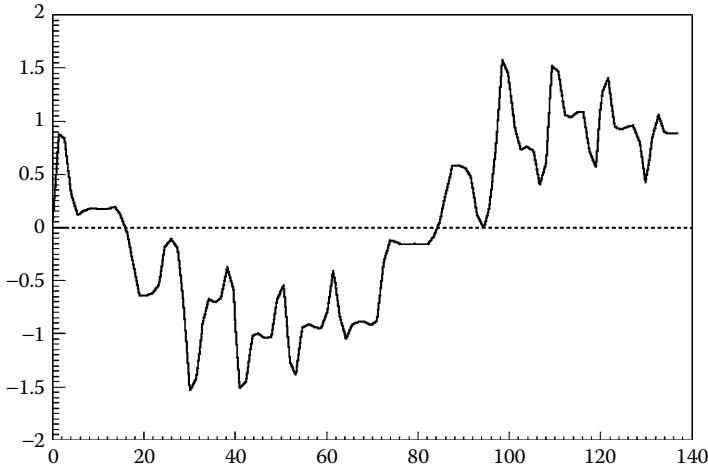


FIGURE 8.8 Permanent magnet (PM) airgap flux density at peak torque.

Now, the flux linkages Ψ_{dm}, Ψ_{qm} are

$$\begin{aligned} \Psi_{dm} &= L_{dm} i_d & L_{dm}(i_d, i_q) &= \frac{\Psi_{dm}}{i_d} \\ \Psi_{qm} &= L_{qm} i_q - \Psi_{PM} & L_{qm}(i_d, i_q) &= \frac{\Psi_{qm} + \Psi_{PMq}}{i_q} \end{aligned} \quad (8.3)$$

Ψ_{PMq} is the flux produced by the PMs in phase A when aligned to rotor axis q , for zero stator currents. This way it is possible to draw, via FEM, a family of curves $\Psi_{dm}(i_{dm}, i_{qm}), \Psi_{qm}(i_{qm}, i_{dm})$ (Figure 8.11a and Figure 8.11b) [9].

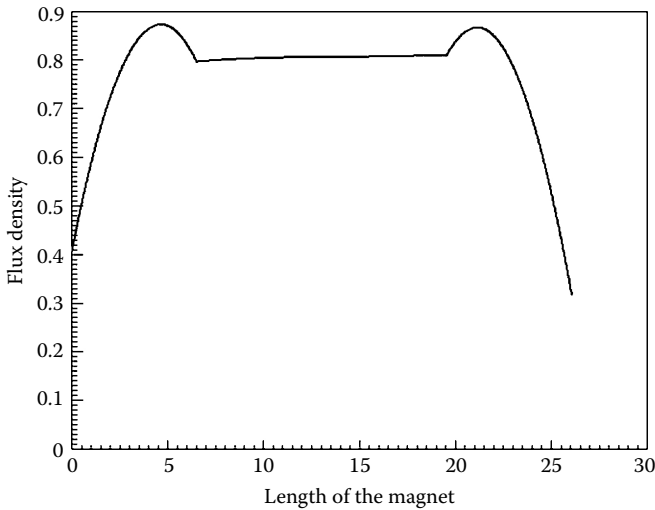


FIGURE 8.9 Permanent magnet (PM) surface flux density at peak torque.

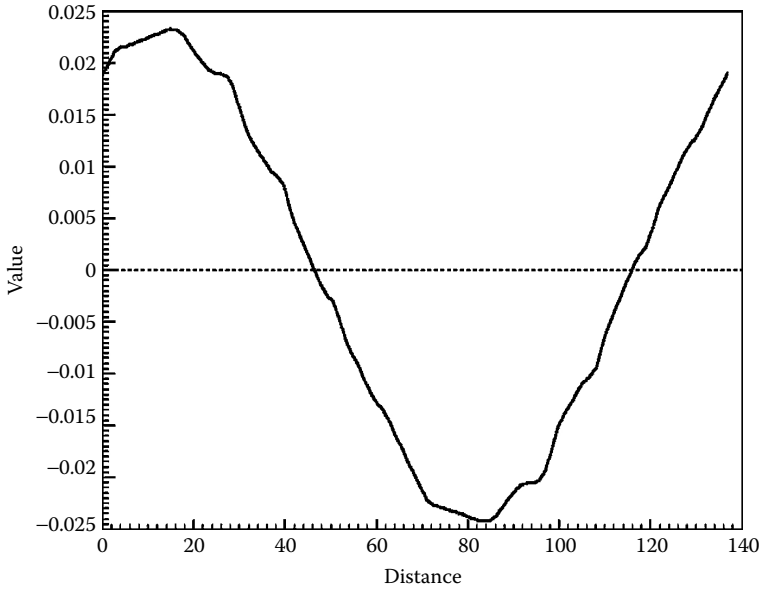


FIGURE 8.10 The airgap flux for the maximum torque.

Figure 8.11a and Figure 8.11b show that the magnetic saturation is notable, especially in axis d , and though it exists, the cross-coupling magnetic saturation is not generally important [10,11] for PM-RSM. For extremely high electric and magnetic loadings with peak current densities of 30 A/mm², the situation might change; thus, direct operation with the family of d - q flux/current curves may prove necessary, via curve fitting.

However, for most cases, $\Psi_d(i_d)$ curve suffices as $L_{qm} = \text{const}$:

$$\begin{aligned} \Psi_{dm} &= L_{dm}(i_d) \cdot i_d & \Psi'_d &= \Psi_{dm} + L_{sl} i_d \\ \Psi_{qm} &= L_{qm}(i_q) \cdot i_q - \Psi_{PMq} & \Psi'_q &= \Psi_{qm} + L_{sl} i_q \end{aligned} \quad (8.4)$$

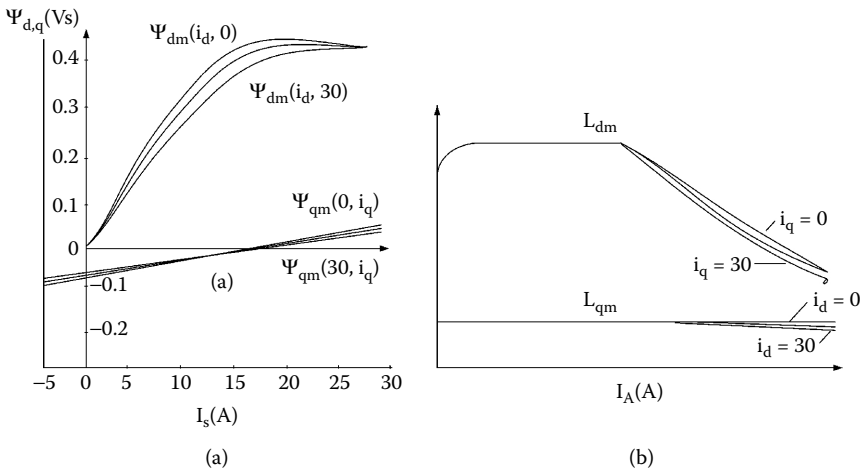


FIGURE 8.11 (a) d - q axes fluxes vs. current families and (b) corresponding L_{dm} and L_{qm} .

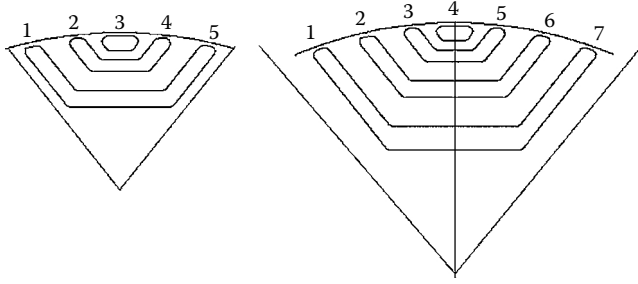


FIGURE 8.12 Rotor poles with 5, 7 flux bridges.

with, for example,

$$L_{dm} = L_{dm0} - \beta(I_d - I_{ds}) - \gamma(I_d - I_{ds})^2 \quad (8.5)$$

Note that the q axis magnetization inductance L_{qmo} may depend on i_q , especially at small values of currents, as the iron bridges above the flux barriers may not be fully saturated at zero stator current. Also, the two-dimensional (2D) FEM analysis might not be satisfactory in predicting L_{qmo} or Ψ_{MPq} , especially in short core stacks when notable axial flux leakage may occur. Three-dimensional (3D) FEM is recommended to avoid the underestimation of L_{qmo} . Also noticed on 2D FEM analyzer is the variation (pulsation) of L_{qmo} with rotor position at slot frequency. This phenomenon is confirmed by tests.

8.3.3 The Cogging Torque

The torque vs. rotor position developed at zero stator current is called the cogging torque. The presence of stator N_s slot openings and the N_r saturated flux bridges in the rotor with PMs placed in the flux barriers, is likely to create the variation of stored magnetic energy in the airgap with rotor position. If we consider N_s and N_r as the number of stator and rotor “poles,” then the cogging torque fundamental number of periods per mechanical revolution $N_{cogging}$ is

$$N_{cogging} = \text{smallest common multiplier of } N_s \text{ and } N_r$$

The higher $N_{cogging}$ is the smaller the cogging torque.

For $q_1 = 2$ and 4 flux barriers per pole in the rotor and $2p_1 = 8$ poles, $N_s = 2p_1q_1m_1 = 2 \cdot 4 \cdot 2 \cdot 3 = 48$ and $N_r = 2 \cdot p_1 \cdot 4 = 32$, $N_{cogging} = 96$, which is a pretty large number. The situation is not practical with the same machine and six flux barriers per pole, however, when $N_r = N_s = 48 = N_{cogging}$. With seven flux bridges per pole, however, $N_r = 7 \cdot 8 = 56$, and $N_{cogging} = 8 \cdot 6 \cdot 7 = 336$, a more favorable design is obtained (Figure 8.12). For the example in this paragraph, the cogging torque obtained through 2D FEM is shown in Figure 8.13.

The very large number of periods in the cogging torque and its rather low value (percent of peak torque of 140 Nm) are evident. Numerous methods to reduce cogging torque in IPM machines — with one PM piece and flux barrier per pole — were proposed. Using FEM to verify them is usual [1], but analytical approaches were also introduced [12] with good results. In essence, the single flux barrier radial side angle and other geometrical parameters of the latter are modified in an orderly manner (direct geometry optimization) to minimize cogging torque while limiting the electromagnetic field (emf) and main torque reductions. Generalizing such methods for multiple flux barrier rotors is more complicated and still to come.

Usual methods to reduce cogging torque, such as stator fractionary windings ($q = 3/2$, for example) or stator slots skewing by up to one slot pitch are also at hand. But, avoiding the undesirable stator slots

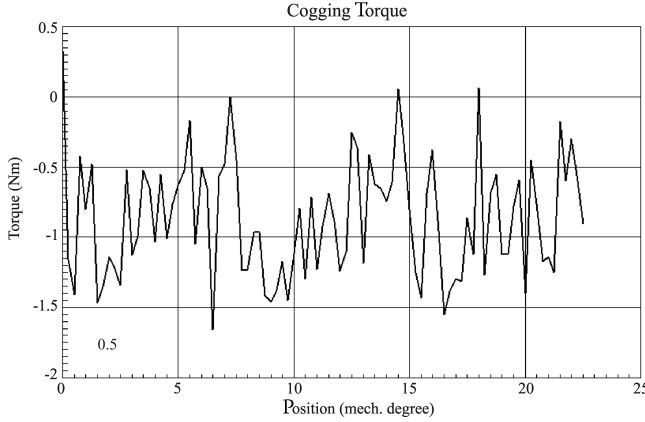


FIGURE 8.13 Cogging torque vs. rotor position.

N_s rotor flux bridges (ribs), N_r , combinations with the largest common multiplier are the modality to avoid torque pulsations (five or seven flux bridges per pole are seen here as practical solutions).

8.3.4 Core Losses Computation by FEM

The advanced magnetic saturation at peak torque, and small speeds, and the lower but harmonic rich flux density in various stator and rotor core zones at high speeds, make the computation of core losses a difficult task. Fundamental and space and harmonics losses are to be considered. Rotor core losses are not to be neglected. On top of that, PM eddy current losses at higher speeds (frequency) are also worthy of consideration. It seems that FEM is indispensable for such a task. In Reference [13], an FEM model for core losses is proposed for the single flux barrier (PM) per pole rotor. Still, the PM eddy current losses are not considered. The procedure could be generalized for the PM-RSM without major obstacles.

In essence, the flux density in the various finite elements is calculated N times for a period, and its radial and tangential components B_r , B_θ are determined. Then, the flux density variations are divided by the time step Δt and used to determine the eddy current and hysteresis losses in the stator and in the rotor [13]:

$$\begin{aligned}
 P_{iron} = & K_{eddy} \cdot \frac{\gamma_{iron}}{2\pi^2} \oint_{iron} \frac{1}{N} \sum_{k=1}^N \left[\left(\frac{B_r^{k+1} - B_r^k}{\Delta t} \right)^2 + \left(\frac{B_\theta^{k+1} - B_\theta^k}{\Delta t} \right)^2 \right] dV \\
 & + K_{hys} \cdot \frac{\gamma_{iron}}{T} \sum_{i=1}^{NE} \frac{\Delta V_i}{2} \left(\sum_{j=1}^{N_{pr}^i} (B_{m\gamma}^{ij})^2 + \sum_{j=1}^{N_{p\theta}^i} (B_{m\theta}^{ij})^2 \right)
 \end{aligned} \tag{8.6}$$

where

γ_{iron} is the specific iron weight

K_{eddy} , K_{hys} are the specific eddy current and hysteresis coefficients, to be determined on the Epstein frame in variable frequency tests, after dividing losses with frequency (Figure 8.14)

N is the number of ΔT time steps for period T

NE is the number of value of elements

ΔV_i is the volume of finite element

B_{mr}^{ij} , $B_{m\theta}^{ij}$ are the amplitudes of each hysteresis loop

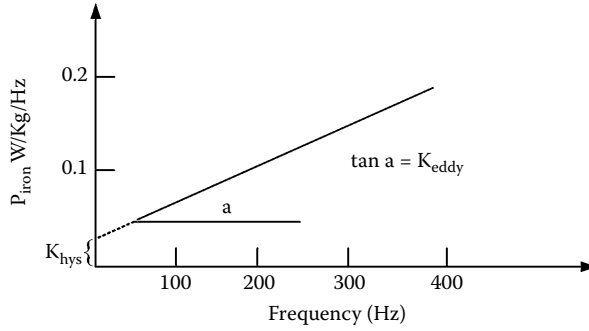


FIGURE 8.14 Iron losses/weight/frequency vs. frequency by the Epstein frame.

The first term in Equation 8.6 represents the eddy current losses, while the second is related to hysteresis losses. The eddy current terms may be extended to the PM body zones with minor adaptations. However, using three to five PM pieces in the axial direction leads to a drastic reduction of PM eddy current losses. Notice that this way, for given stator current amplitude, current power angle δ_{idb} and speed, the period of the current is explored for the fundamental. However, the current time harmonics also have to be considered.

The amount of computation time to explore core losses for the entire spectrum of torque and speed, for motoring and generating, becomes large. An equivalent core resistance R_C , even if slightly variable with frequency, may be defined:

$$\frac{1}{R_C} = \frac{1}{R_{eddy}} + \frac{1}{\omega R_{hys}} \quad (8.7)$$

R_{eddy} corresponds to eddy current and is constant. A frequency-dependent term (different in the stator than in the rotor) has to be added, together with a constant R_{hys} term, to represent the hysteresis losses in the d - q model. The term R_C acts in parallel to the total stator flux (emf) to represent the core losses:

$$R_C = \frac{3}{2} \frac{(\omega_1 \Psi_s)^2}{P_{iron}} \quad (8.8)$$

in the d - q model.

A typical breakdown of iron losses at two different speeds for a 2.5 Nm peak torque, four-poles IPM machine (with one flux barrier per pole) is shown in Figure 8.15 [13]. It should be noticed that the stator current produces notable rotor iron losses due to stator-mmf and slot-opening space harmonics. For higher speeds, as expected in starter/alternators, rotor core losses become important. Good efficiency correlation up to 6000 rpm (200 Hz) is claimed with this method in Reference [13].

8.4 The d - q Model of PM-RSM

As inferred from FEA, in a professionally designed PM-RSM:

- Magnetic saturation manifests itself along the d and q axes separately, and there is more along the d axis.
- Despite magnetic saturation, the PM and total flux in stator phases with distributed windings varies sinusoidally in time, for sinusoidal currents and constant speed.

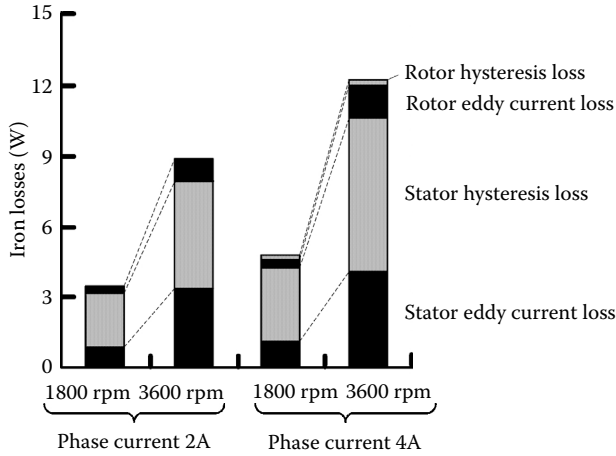


FIGURE 8.15 Calculated iron losses, via finite element method (FEM).

- Stator and rotor core losses may be represented through a core resistance R_C that is slightly dependent on frequency and “acts” in parallel to the total (stator) flux produced emf: $\omega_1 \Psi_s$.
- The damping effect of rotor core losses is neglected. In such conditions, the d - q model can be applied directly to the PM-RSM. First, the core losses are neglected, for simplicity, and then they are added to the model for completeness. As PM-RSM is, in fact, a salient-pole synchronous machine, the d - q model has to be attached to the rotor reference system [4]:

$$\bar{V}_s = R_s \bar{I}_s + j\omega_r \bar{\Psi}_s + \frac{d\bar{\Psi}_s}{dt} \quad (8.9)$$

$$\bar{\Psi}_s = \Psi_d + j\Psi_q$$

$$\Psi_d = L_d(i_d) \cdot i_d \quad (8.10)$$

$$\Psi_q = L_q i_q - \Psi_{PMq}$$

$$T_e = \frac{3}{2} p (\Psi_d i_q - \Psi_q i_d) = \frac{3}{2} p (\Psi_{PMq} + (L_d - L_q) i_q) i_d \quad (8.11)$$

$$\frac{J}{p} \frac{d\omega_r}{dt} = T_e - T_{load} \quad (8.12)$$

Equation 8.9 through Equation 8.12 are written for the motoring mode with positive currents and powers. For the generator mode, the torque has to be negative, and thus, i_d has to change sign. Also, the T_{load} becomes negative, as it turns into the negative driving torque of the prime mover. The active power is negative for generating in Equation 8.9 through Equation 8.12.

The relationships between stator phase quantities and the d - q model variables are, as known [4],

$$\bar{V}_s = \frac{2}{3} \left(V_a + V_b e^{j\frac{2\pi}{3}} + V_c e^{-j\frac{2\pi}{3}} \right) e^{-j\theta_{er}}; \quad \frac{d\theta_{er}}{dt} = \omega_r \quad (8.13)$$

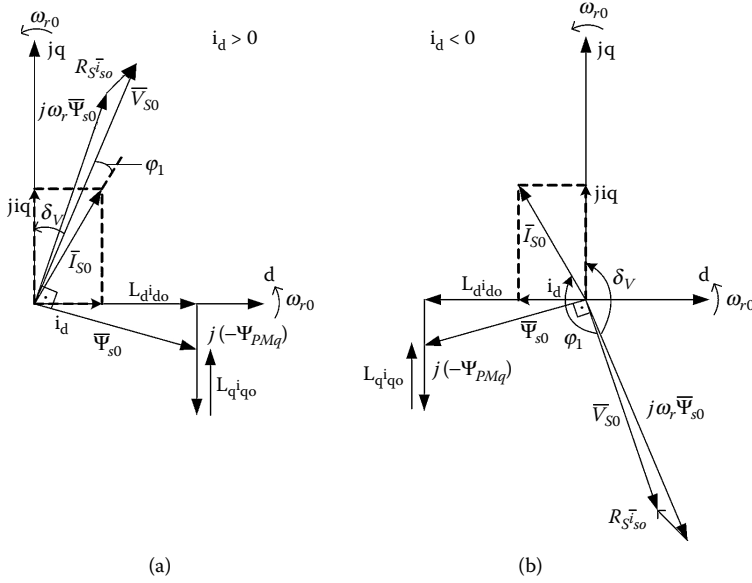


FIGURE 8.16 Vector diagrams of permanent-magnet-assisted reluctance synchronous machine (PM-RSM): (a) motoring and (b) generating.

The same transformation is valid for currents and flux linkages. For steady state, with sinusoidal phase voltages of frequency $\omega_1 = \omega_{r0}$,

$$V_{a,b,c} = V\sqrt{2} \cos\left((\omega_{r0}t + \gamma_o) - (i-1)\frac{2\pi}{3}\right) \quad (8.14)$$

$$\bar{V}_{s0} = V\sqrt{2}(\cos\gamma_o - j\sin\gamma_o) \quad (8.15)$$

Consequently, steady state, in rotor coordinates, corresponds to direct current (DC) voltages, currents, and flux linkages. So, $d/dt = 0$ for steady state in Equation 8.9, and $d\omega_r/dt = 0$ in Equation 8.12.

The vector diagrams for motoring ($i_d > 0$) and generating ($i_d < 0$) are shown in Figure 8.16a and Figure 8.16b, respectively, for the trigonometrical positive direction of motion. The flux vector is behind stator current for motoring and ahead for generating. From the vector diagrams,

$$i_d = I_s \sin(\delta_V + \phi_1) \quad (8.16)$$

$$i_q = I_s \cos(\delta_V + \phi_1) \quad (8.17)$$

It should be noted that the q axis current is demagnetizing (defluxing) the machine, while i_d is the main torque current now. From the vector diagram,

$$V_d + jV_q = \omega_r(L_{q0}i_{q0} - \Psi_{PMq}) - R_s i_{d0} + j(R_s i_{q0} + \omega_r L_d(i_d)i_d) \quad (8.18)$$

$$V_s = \sqrt{V_d^2 + V_q^2}$$

$$V_s^2 = (\omega_r(L_{q0}i_{q0} - \Psi_{PMq}) - R_s i_{d0})^2 + (R_s i_{q0} + \omega_r L_d(i_d)i_d)^2 \quad (8.19)$$

Again, $i_{d0} > 0$ for motoring and $i_{d0} < 0$ for generating and $i_{q0} > 0$ for both operation modes, in general.

For given torque T_{e0} , stator vector voltage V_{s0} , and speed, Equation 8.19 and Equation 8.11 provide the values of i_d and i_q . An iterative procedure is required because of the nonlinearity of the equations and due to magnetic saturation ($L_d(i_{d0})$).

The efficiency and power factor may now be calculated for given torque, speed, and stator voltage:

$$\eta_e = \frac{T_e \omega_r / p}{T_e \omega_r / p + \frac{3}{2} R_s (i_{d0}^2 + i_{q0}^2)}, \quad \text{for motoring} \quad (8.20)$$

$$\cos \phi_1 = \frac{|T_e \omega_r / p|}{\eta_e \cdot \frac{3}{2} V_{s0} \sqrt{i_{d0}^2 + i_{q0}^2}} \quad (8.21)$$

Note that only the stator winding losses have been considered in Equation 8.20 and Equation 8.21. Finally, from Equation 8.16, the voltage power angle may be obtained. Alternatively, the steady-state performance calculation may start with given V_{s0} , ω_r , δ_v :

$$\begin{aligned} V_{d0} &= V_{s0} \sin \delta_v & 90^\circ > \delta_v > 0 & \text{for motoring} \\ V_{q0} &= V_{s0} \cos \delta_v & 90^\circ < \delta_v & \text{for generating} \end{aligned} \quad (8.22)$$

Then,

$$\begin{aligned} V_{d0} &= \omega_r (L_{q0} i_{q0} - \Psi_{PMq}) - R_s i_{d0} \\ V_{q0} &= \omega_r L_d (i_{d0}) i_{d0} + R_s i_{q0} \end{aligned} \quad (8.23)$$

With the magnetic saturation level defined ($L_d = \text{constant}$), Equation 8.22 and Equation 8.23 yield the currents i_{d0} and i_{q0} analytically. Then the torque, efficiency, and power factor may all be computed from Equation 8.11 and Equation 8.20 and Equation 8.21. If the stator resistance R_s is neglected,

$$i_{d0} = \frac{V_{s0} \cos \delta_v}{\omega_r L_d}; \quad \delta_v > 90^\circ \text{ for generating} \quad (8.24)$$

$$i_{q0} = \frac{V_{s0} \sin \delta_v + \omega_r \Psi_{PMq}}{\omega_r L_q} \quad (8.25)$$

$$(\cos \phi_1)_{R_s=0} = \frac{\left| \frac{3}{2} (\Psi_{PMq} + (L_d - L_q) i_{q0}) i_{d0} \right| \omega_r}{\frac{3}{2} V_{s0} \sqrt{i_{d0}^2 + i_{q0}^2}} \quad (8.26)$$

Example 8.1

Consider a PM-RSM with the constant parameters $L_d = 200$ mH, $L_q = 60$ mH, $\Psi_{PMq} = 0.215$ Wb, $2p_1 = 4$ poles, rated voltage 220 V per phase. Neglecting the stator resistance, calculate the steady-state torque, current, and power factor vs. voltage power angle δ_v for motoring at 1500 rpm and a stator voltage of 220 V (RMS). Increase PM flux twice, and check the performance at $\delta_v = 60^\circ$.

Solution

The voltage vector amplitude V_{s0} is (Equation 8.15):

$$V_{s0} = V \sqrt{2} = 220 \sqrt{2} = 310.2 \text{ V}$$

TABLE 8.2 PM-RSM Characteristics

Voltage Power Angle $\nu(^{\circ})$	I_{do} (A)	I_{qo} (A)	I_{so} (A)	T_{eo} (nm)	$\cos_1$
0	0	20.048	20.048	0	0
15	1.278	19.48	19.5	11.29	0.1942
30	2.469	17.858	18.028	20.28	0.37744
45	3.4922	15.22	15.62	24.58	0.5276
60	4.2825	11.816	12.498	24.01	0.678
75	4.77	7.844	9.18	18.79	0.686
90	4.939	3.583	6.10	10.619	0.5837

Now, we just follow Equation 8.24 through Equation 8.26 for a few values of δ_v , and we obtain the results in Table 8.2.

Notice that the emf (PM-induced voltage) $E_{s0} = 67.51$ V, while $V_{s0} = 310.2$ V. This means that the PM contribution is, in such conditions, rather small. As the saliency ratio $L_d/L_q = 200/60 = 3.33$ only, it is no wonder that the power factor does not rise above 0.686. The maximum torque is moved from the 45° that characterizes the purely RSM to somewhere between 45° and 60° . With a stronger (doubled) PM contribution, the torque is increased at $\delta_v = 60^{\circ}$ from 24.01 Nm to 33.22 Nm, while the power factor is only slightly improved as, still, $E_{s0}/V_{s0} = 135$ V/310.2 V.

When the speed increases further, the power factor is likely to be notably increased for constant (given) voltage, as E_{s0} approaches or even surpasses V_{s0} by, in general, at most 50% at highest admissible speed, to avoid voltage oversizing of pulse-width modulated (PWM) converter power switches.

The torque vs. voltage power angle is shown in Figure 8.17, for zero losses, constant voltage, and speed.

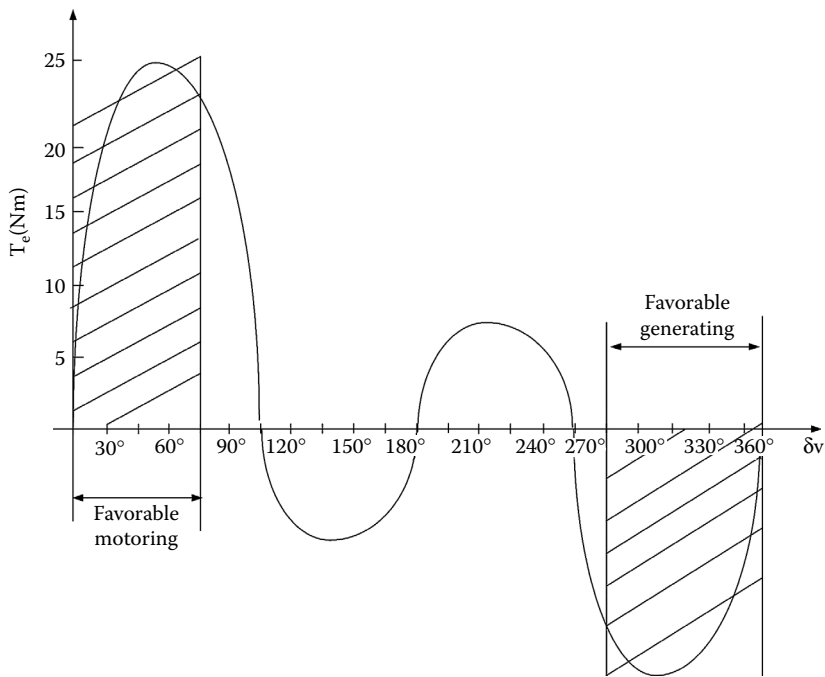


FIGURE 8.17 Typical permanent-magnet-assisted reluctance synchronous machine (PM-RSM) torque vs. voltage power angle curve for constant voltage and speed.

The expression of torque vs. voltage power angle δ_v , for zero stator resistance, is derived simply from Equation 8.24 through Equation 8.26:

$$T_e = \frac{3}{4} p \left(\frac{V_{s0}}{\omega_{r0}} \right)^2 \left(\frac{1}{L_q} - \frac{1}{L_d} \right) \sin 2\delta_v + \frac{3}{2} p \frac{V_{s0}}{\omega_{r0}} \frac{\Psi_{PMq}}{L_q} \sin \delta_v \quad (8.27)$$

The constant voltage performance is important during motoring and generating inside the constant power speed range, where the PWM converter has to operate at full voltage. However, operation at constant voltage for a wide range of speeds encounters widely variable magnetic saturation and current conditions; thus, it has to be carefully watched through adequate control. Also, the current should be limited to the peak design value. At low speeds, the stator resistance may not be neglected. The voltage power angle should vary with speed to produce the required torque for minimum current up to base speed ω_b and for minimum flux Ψ_s at high speeds. The d - q model equations may be written for transients as follows:

$$V_d = R_s i_d + \frac{d}{dt} (L_d(i_d) i_d) - \omega_r (L_q i_q - \Psi_{PMq}) \quad (8.28)$$

$$V_q = R_s i_q + \frac{d}{dt} (L_q i_q) + \omega_r L_d(i_d) i_d \quad (8.29)$$

Also,

$$\frac{d}{dt} (L_d(i_d) i_d) = \left(L_d(i_d) + i_d \frac{dL_d}{di_d} \right) \frac{di_d}{dt} = L_d^t(i_d) \frac{di_d}{dt} \quad (8.30)$$

$L_d^t(i_d)$ is the transient inductance along axis d , while $L_d(i_d)$ is the normal (rotational) inductance. In general, $L_d^t < L_d$ for saturated core conditions, and $L_d^t = L_d$ under unsaturated conditions. At very low values of current, again, $L_d^t < L_d$ due to the early inflexion in the core magnetization curve.

In case the magnetic saturation along axis q is also considered,

$$\frac{d}{dt} (L_q i_q) = \left(L_q + i_q \frac{dL_q}{di_q} \right) \frac{di_q}{dt} = L_q^t(i_q) \frac{di_q}{dt} \quad (8.31)$$

Equation 8.28 through Equation 8.41 lead to the general equivalent circuits for transients shown in [Figure 8.18a](#) for axis d and [Figure 8.18b](#) for axis q .

The core losses resistance R_C is “planted across the motion induced emf” and strictly considers the core loss in the stator during steady state. To account for rotor core and PM eddy current losses during transients or due to harmonics, a special additional circuit may be added in parallel to L_{ds}^t and L_{qs}^t terms, as a kind of weak damper case in the rotor ([Figure 8.18a](#) and [Figure 18.8b](#)).

The equivalent damper winding parameters may be determined from standstill DC current decay or frequency tests as done for synchronous machines. For transients, R_C may be neglected ($R_C = \infty$). The addition of the core loss resistance R_C , in the absence of the damper cages ($R_D = R_Q = 0$, $L_{Dl} = L_{Ql} = 0$), changes, to some extent, the equations of the d - q model to the following:

$$V_d = R_s i_d + L_d^t \frac{di_d}{dt} - \omega_r (L_q i_q - \Psi_{PMq}) \quad (8.32)$$

$$V_q = R_s i_q + L_q^t \frac{di_q}{dt} + \omega_r L_d i_d \quad (8.33)$$

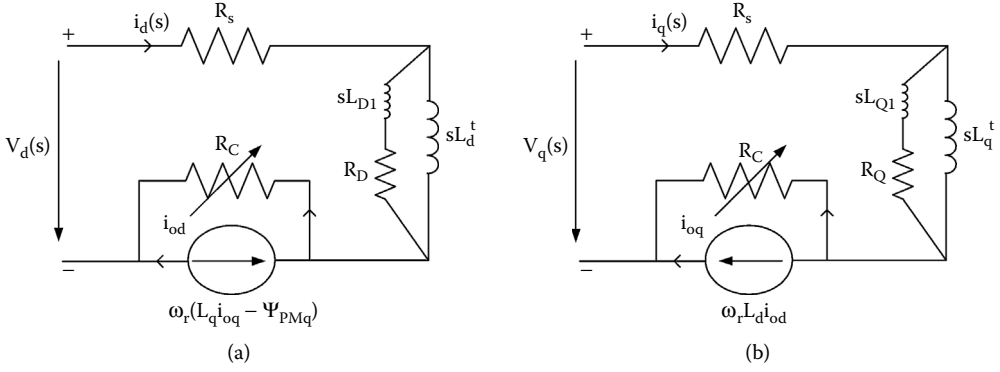


FIGURE 8.18 Permanent-magnet-assisted reluctance synchronous machine (PM-RSM) equivalent circuits for transients (a) for axis d and (b) for axis q .

$$i_{0d} = i_d + \frac{\omega_r (L_q i_{0q} - \Psi_{PMq})}{R_C} \quad (8.34)$$

$$i_{0q} = i_q - \frac{\omega_r L_d i_{0d}}{R_C} \quad (8.35)$$

The torque is as follows:

$$T_e = \frac{3}{2} p (\Psi_{PMq} + (L_d - L_q) i_{0q}) i_{0d} \quad (8.36)$$

Again, for steady state $\frac{d}{dt} = 0$. The core and windings losses are as follows:

$$P_{iron} = \frac{3}{2} \frac{\omega_r [(L_q i_{0q} - \Psi_{PMq})^2 + L_d^2 i_{0d}^2]}{R_C} \quad (8.37)$$

$$P_{copper} = \frac{3}{2} R_s (i_d^2 + i_q^2) \quad (8.38)$$

The efficiency η and power factor $\cos \phi_1$ become

$$\eta = \frac{\frac{3}{2} (V_d i_d + V_q i_q) - P_{iron} - P_{copper} - P_{mec}}{\frac{3}{2} (V_d i_d + V_q i_q)} \quad (8.39)$$

$$\cos \phi_1 = \frac{(V_d i_d + V_q i_q)}{\sqrt{V_d^2 + V_q^2} \cdot \sqrt{i_d^2 + i_q^2}} \quad (8.40)$$

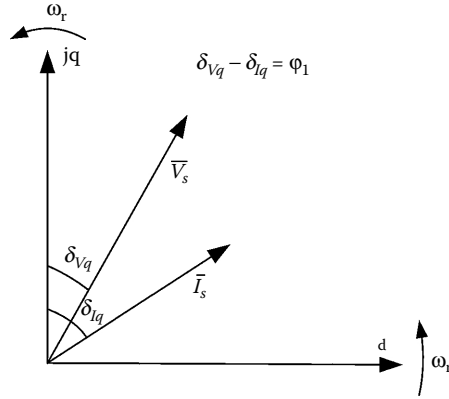


FIGURE 8.19 Control power angles to assess constant voltage (or constant current) and speed performance.

P_{mec} are the mechanical losses. Additional losses are lumped either into the copper or into core losses. The complete model is to be used for realistic performance computation of PM-RSM.

Once accepting this, we run into the difficulty of accounting for magnetic saturation functions $L_d(i_d)$ and $L_q(i_q)$ based on which $L_d^*(i_d)$, $L_q^*(i_q)$ may be calculated.

For steady state, only the $L_d(i_d)$ and $L_q(i_q)$ functions are needed, and a simple iterative computation procedure may be developed to extract the performance for given current i_{s0} , ω_r for various values of the current power angle $\delta_{i_q} = \tan^{-1} I_d/I_q$; or for given voltage V_{s0} and ω_r for various values of the voltage power angle $\delta_{V_q} = \tan^{-1} V_q/V_d$ ($\delta_V = \delta_{V_q}$ as defined previously in this paragraph; Figure 8.19).

8.5 Steady-State Operation at No Load and Symmetric Short-Circuit

8.5.1 Generator No-Load

Generator no-load operation ($i_d = i_{q0} = 0$) at steady state, Equation 8.32 through Equation 8.35, becomes as follows:

$$i_{0d} = \frac{\omega_r (L_q i_{q0} - \Psi_{PMq})}{R_C} \quad (8.41)$$

$$i_{0q} = -\frac{\omega_r L_d i_{0d}}{R_C} \quad (8.42)$$

So,

$$i_{0d} = -\frac{\Psi_{PMq} \omega_r \cdot R_C}{R_C^2 + \omega_r^2 L_d I_q} < 0; \quad i_{0q} > 0 \quad (8.43)$$

$$V_{d00} = -\omega_r (L_q i_{q0} - \Psi_{PMq}) \quad (8.44)$$

$$V_{q00} = \omega_r L_d i_{0d} \quad (8.45)$$

when $R_C = \infty$ (zero iron losses), $V_{d00} = +\omega_r \Psi_{PMq}$, $V_{q00} = 0$, and $i_{0d} = i_{0q} = 0$, as expected.

8.5.2 Symmetrical Short-Circuit

For steady-state symmetrical short-circuit $V_d = V_q = 0$, and thus,

$$0 = R_s \left(i'_{0d} - \frac{\omega_r (L_q i'_{0q} - \Psi_{PMq})}{R_C} \right) - \omega_r (L_q i'_{0q} - \Psi_{PMq}) \quad (8.46)$$

$$0 = R_s \left(i'_{0q} + \frac{\omega_r L_d i'_{0d}}{R_C} \right) + \omega_r L_d i'_{0d} \quad (8.47)$$

$$i_{dsc} = i'_{0d} - \frac{\omega_r L_{0q} i_{0q} - \Psi_{PMq}}{R_C} \quad (8.48)$$

$$i_{qsc} = i'_{0q} + \frac{\omega_r L_d i'_{0d}}{R_C} \quad (8.49)$$

Especially at high speeds (frequencies), the core losses are notable and may not be neglected. The torque at symmetrical steady-state short-circuit may still be calculated by using Equation 8.36. With i'_{0d} and i'_{0q} easy to extract from Equation 8.46 and Equation 8.47, i_{dsc} and i_{qsc} are straightforwardly obtained from Equation 8.48 and Equation 8.49. Equation 8.46 and Equation 8.47 may be written in the following form:

$$R_s i'_{0d} - \omega_r L_q \left(1 + \frac{R_s}{R_C} \right) i'_{0q} = -\omega_r \Psi_{PMq} \left(1 + \frac{R_s}{R_C} \right) \quad (8.50)$$

$$\omega_r L_d \left(1 + \frac{R_s}{R_C} \right) i'_{0d} + R_s i_{0q} = 0 \quad (8.51)$$

Solving for i'_{0q} and i'_{0d} in Equation 8.50 is again straightforward. The electromagnetic power $\frac{T_{esc} \omega_r}{P_l}$ is transformed into losses of copper and iron. When these losses are neglected ($R_s = 0$, $R_C = \infty$),

$$i'_{0d} = i_{dsc} = 0, i'_{0q} = \frac{\Psi_{PMq}}{L_q} = i_{qsc} = i_{sc3i}$$

So, the ideal short-circuit current is proportional to the PM flux linkage Ψ_{PMq} and inversely proportional to L_q ($L_q < L_d$). The higher the Ψ_{PMq} and the lower L_q (higher saliency L_d/L_q), the higher the short-circuit current. In most wide constant power speed range design,

$$\Psi_{PMq} \approx L_q i_{srated} \quad (8.52)$$

For ideal no-load motoring — zero electromagnetic torque — and zero mechanical losses, from Equation 8.36, $i_{0d} = 0$, and from Equation 8.32 and Equation 8.33 at steady state ($\frac{d}{dt} = 0$),

$$i_d = -\frac{\omega_r (L_q i_{0q} - \Psi_{PMq})}{R_C} \quad (8.53)$$

$$V_d = -\frac{R_s}{R_C} \omega_r (L_q i_{0q} - \Psi_{PMq}) - \omega_r (L_q i_{0q} - \Psi_{PMq})$$

$$V_q = R_s i_{0q}; \quad i_q = i_{0q} \quad (8.54)$$

$$V_s^2 = V_d^2 + V_q^2$$

The three expressions in Equation 8.54 serve to calculate first i_{0q} and then V_d and V_q , and finally, with Equation 8.53, i_d and i_{q0} .

Example 8.2

For the PM-RSM in Example 8.1, calculate the no-load generator voltage V_{s0} at 3000 rpm and the short-circuit symmetric current at the same speed for zero losses in the machine.

Solution

For the generator under no load, the terminal voltage V_{d00} (Equation 8.44) is as follows:

$$V_{d00} = E_{s0} = \omega_r \Psi_{PMq} = 2\pi \frac{3000 \times 2}{60} \times 0.215 = 135 \text{ V (peak voltage per phase)}$$

For the same speed, the ideal symmetrical short-circuit $i_{qsc} = i_{sc3}$ (Equation 8.51) is as follows:

$$i_{sc3} = \frac{\Psi_{PMq}}{L_q} = \frac{0.215}{0.06} = 3.5833 \text{ A}$$

With zero losses, the short-circuit torque is zero.

8.6 Design Aspects for Wide Speed Range Constant Power Operation

In automotive applications, a wide speed range constant power operation is required. Speed range (ω_{max}/ω_b) up to 10 (12) would be desirable mostly for generating, but also for motoring, in the same cases. The typical torque/speed and voltage/speed envelopes are given in Figure 8.20.

The base speed ω_b is defined as the maximum speed for which the peak (starting) torque is produced at maximum converter voltage V_{smax} under the condition of maximum torque/current criterion. With wide speed range for constant power ($F > 2$ to 2.5), oversizing of either the PM-RSM or of the PWM converter or of both is required.

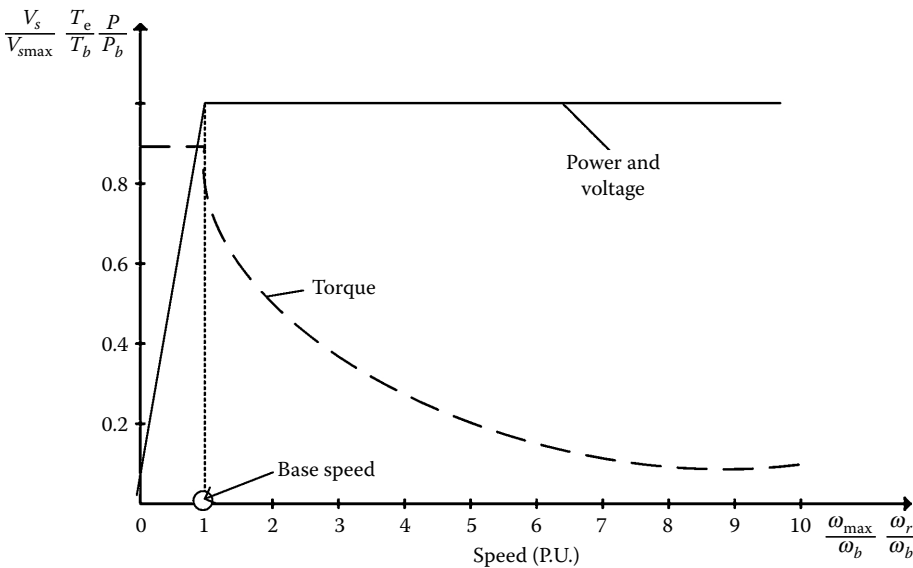


FIGURE 8.20 Torque, power, and voltage vs. speed.

In general, machine oversizing (overrating) is measured in over-torque, while converter oversizing is expressed in current oversizing at the maximum ideal converter line natural peak voltage fundamental, V_{smax} , given for all designs:

$$V_b = V_{smax} = \frac{2}{\pi} V_{dc} \quad (8.55)$$

To investigate the oversizing aspects, a normalization system provides the required degree of generality. The base electrical quantities are as follows:

- V_b is the maximum machine voltage
- I_b is the maximum machine phase current
- ω_b is the base speed that corresponds to the speed at which the machine first reaches V_b at I_b for maximum torque/current criterion
- T_b is the base torque that is produced at I_b for maximum torque/current criterion
- Ψ_b is the base flux: $\Psi_b = \frac{V_b}{\omega_b}$

To simplify the analysis, let us neglect losses in the machine at steady state. From Equation 8.9 through Equation 8.12,

$$\begin{aligned} \bar{V}_{s0} &= j\omega_b \bar{\Psi}_{s0} \\ \bar{\Psi}_{s0} &= L_d \bar{i}_d + j(L_q \bar{i}_q - \Psi_{PMq}) \\ I_b &= \sqrt{\bar{i}_d^2 + \bar{i}_q^2} \\ T_e &= \frac{3}{2} p_1 (\Psi_{PMq} + (L_d - L_q) \bar{i}_q) \bar{i}_d \end{aligned} \quad (8.56)$$

Let us neglect magnetic saturation variation and consider $L_d = \text{constant}$. The maximum torque for given current I_b is explored in what follows.

The condition $\frac{dT_e}{d\bar{i}_q} = 0$ for given stator current I_b , yields the following:

$$2(L_d - L_q) \bar{i}_q^2 + \bar{i}_q \Psi_{PMq} - (L_d - L_q) I_b^2 = 0 \quad (8.57)$$

$$I_{qb} = \frac{-\Psi_{PMq} + \sqrt{\Psi_{PMq}^2 + 8(L_d - L_q)^2 I_b^2}}{4(L_d - L_q)} \quad (8.58)$$

$$I_{db} = \sqrt{I_b^2 - I_{qb}^2} \quad (8.59)$$

Finally,

$$T_b = \frac{3}{2} p_1 (\Psi_{PMq} + (L_d - L_q) I_{qb}) I_{db} \quad (8.60)$$

Also, the (T_e/I_s) function may be maximized when the maximum Newton meter per ampere of current is obtained.

Note that when magnetic saturation is considered, the above solution does not hold true. I_{qb} tends to be higher than in Equation 8.58. It should be noticed that the rated torque T_n would correspond to V_b and I_b and the unity power factor:

$$T_n = \frac{3}{2} \frac{p}{\omega_b} V_b I_b \neq T_b \quad (8.61)$$

The base and rated torque are different in this normalization, as the power factor of the machine is not unity in general. So, $T_b/T_n < 1$. The base speed ω_b , for given V_b and T_b (i_{db} and i_{qb}) has to be calculated from Equation 8.56:

$$\omega_b = \frac{V_b}{\sqrt{(L_d I_{db})^2 + (L_q I_{qb} - \Psi_{PMq})^2}} \quad (8.62)$$

For the maximum speed at constant power ω_{max} and V_b , the machine is to reach the condition of maximum torque per given flux Ψ_{smin} :

$$\Psi_{smin} \approx \frac{V_b}{\omega_{max}} = \frac{V_b}{\omega_b \cdot F} = \frac{\Psi_b}{F} \quad (8.63)$$

So, from Equation 8.56, with $\frac{dT_e}{di_q} = 0$:

$$\Psi_{smin} = \sqrt{(L_d i_d)^2 + (L_q i_q - \Psi_{PMq})^2} \quad (8.64)$$

In general, magnetic saturation is negligible at maximum speed, so finally,

$$2(L_d - L_q)L_q^2 i_q^2 - (3L_d - 4L_q)L_q \Psi_{PMq} i_q - (L_d - L_q)\Psi_s^2 - L_d \Psi_{PMq}^2 = 0 \quad (8.65)$$

$$(i_q)_{\left(\omega_{smin}^{\max}\right)} = \frac{(3L_d - 4L_q)L_q \Psi_{PMq} + \sqrt{(3L_d - 4L_q)L_q^2 \Psi_{PMq}^2 + 8(L_d - L_q)L_q^2 \cdot [(L_d - L_q)\Psi_{smax}^2 + L_d \Psi_{PMq}^2]}}{4(L_d - L_q)L_q^2} \quad (8.66)$$

Knowing ω_{max} , V_b , that is, Ψ_{smin} , the $(i_q)_{\omega_{max}}$, $(i_d)_{\omega_{max}}$ pair is obtained. Then, the corresponding torque is calculated. Finally, the stator current maximum speed $\omega_{max}(I_s)$ is obtained. If $(T_e)^* \omega_{max}/p \geq P_b$ and $(I_s)\omega_{max} \leq I_b$, then the machine is capable of handling the respective constant speed power range from ω_b to ω_{max} ; if not, oversizing is required. A complete discussion on machine and inverter oversizing for wide F range is given in Reference [14]. So far, we did not mention any design constraint for wide constant power speed range. Reference [15] contains such a thorough first analysis. The main conclusion is that the best designs, if not otherwise constrained, for wide constant power speed range, are obtained if the ideal no-load motoring (zero torque) speed ω_0 is infinite:

$$(\omega_0)_{i_d=0} \cong \frac{V_b}{L_q i_b - \Psi_{PMq}} \cong \infty \quad (8.67)$$

So,

$$\Psi_{PMq} = L_q I_b \quad (8.68)$$

In per unit (P.U.), it means that $\Psi_{PMq} = l_q$. In other words, the flux in the machine at zero torque should be zero. As part of L_q is the leakage inductance, no apparent danger of PM demagnetization occurs from this mathematical cancellation.

It was proven [14] that Ψ_{PMq} (P.U.) and the L_d/L_q ratio remain the independent parameters in the above-presented normalization system. Typical maximum ideal power (zero losses) vs. speed for $L_d/L_q = \zeta$ and three different Ψ_{PMq} (P.U.) and Ψ_{PMq}/L_q ratios are shown in Figure 8.21 [14].

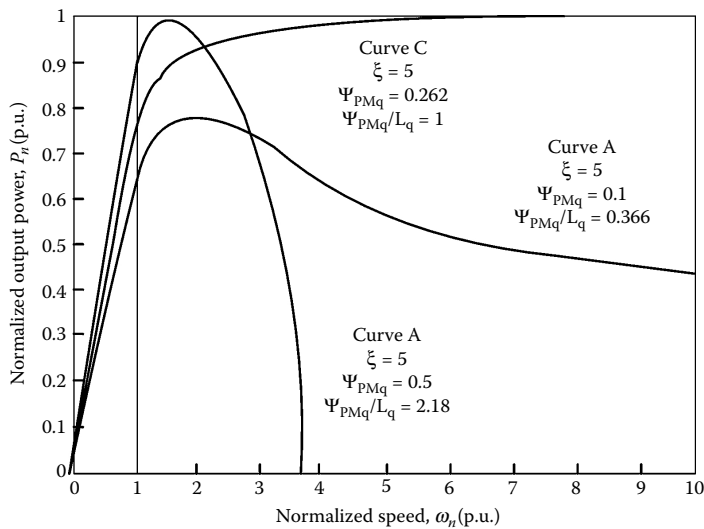


FIGURE 8.21 Maximum power vs. speed for three permanent-magnet-assisted reluctance synchronous machines (PM-RSMs).

The maximum power is obtained at maximum (I_b) current up to base speed ω_b and at I_b current values (or smaller) above ω_b for base voltage V_b . Figure 8.21 contains fairly general results that warrant the following remarks:

- For $\Psi_{PMq}/L_q = 1$ (P.U.), the P.U. power increases steadily toward unity at infinite speed.
- For $\Psi_{PMq}/L_q = 1$, at base speed, the power in P.U. is less than the base power.
- For Ψ_{PMq} large and $\Psi_{PMq}/L_q > 1$, the power peaks quickly and drops rapidly with speed but is large at base speed.
- Even designs with small Ψ_{PMq} and, thus, $\Psi_{PMq}/L_q < 1$ are poor for wide speed ranges, as the level of constant power that can be maintained is smaller.
- A near optimal design for $F = \omega_{max}/\omega = 10/1$ should depart somewhat from the $\Psi_{PMq}/L_q = 1$ condition to yield more torque at base speed (Figure 8.22) for a higher constant power level [14].

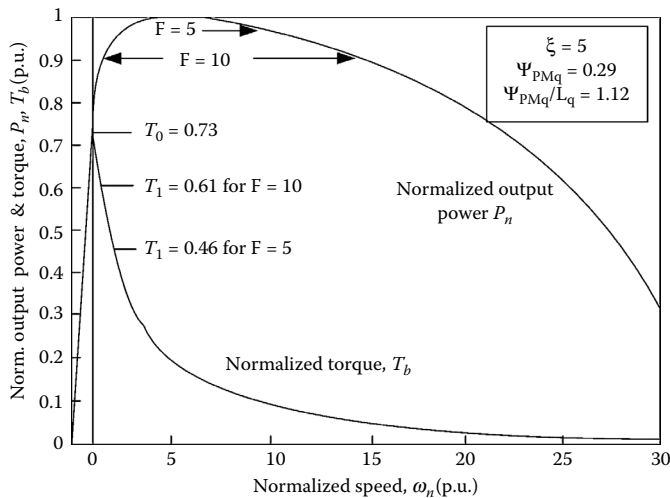


FIGURE 8.22 Near-optimal design for constant power range.

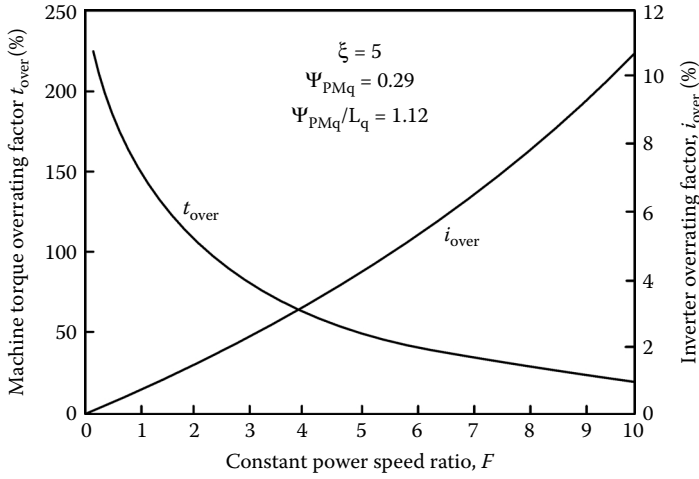


FIGURE 8.23 Oversizing of machine (t_{over}) and of inverter (i_{over}).

- For $F = 10$, ω_{min} and ω_{max} are left to float; and thus, $\omega_{min} > \omega_b$ allows for more power at minimum speed. At $\omega_{min} > \omega_b$ the inverter already works at maximum voltage. This leads to machine oversizing. The machine oversizing is defined in torque terms (Figure 8.22):

$$t_{over} = \left(\frac{T_b}{T_1} - 1 \right) \times 100\% \quad (8.69)$$

The inverter current oversizing i_{over} is as follows:

$$i_{over} = \left(\frac{1}{T_1} - 1 \right) \times 100\% \quad (8.70)$$

For the design in Figure 8.22, for $F = 10$ $t_{over} = 20\%$, $i_{over} = 11\%$. A complete illustration of t_{over} and I_{over} for $L_d/L_q = 5$, $\Psi_{PMq}/L_q = 1.12$ is shown in Figure 8.23 [14].

Smaller inverter oversizing and a larger machine oversizing are preferred when system cost is paramount. Machine volume constraints under the vehicle hood might reduce the advantage of such a choice. A higher speed machine with a transmission might do it in such a case. Though instrumental in securing wide constant power speed range, choosing $\omega_{min} > \omega_b$ leads to a design that reaches full voltage in the midst of its constant torque range. It implies saturation of the current regulators and, thus, impedes on torque control. In contrast, choosing $\omega_{min}/\omega_b < 1$, it means that $P_{\omega_{max}} > (P_1)_{\omega_{min}}$. This time, the inverter oversizing is $i_{over} = \left(\frac{1}{(T_1)_{\omega_{min}}} - 1 \right) \times 100\%$. The machine is now fully utilized ($t_{over} = 0$), and the current regulators will not saturate. The price to pay for full machine utilization is a higher inverter oversizing ($i_{over} = \left(\frac{1}{T_1} - 1 \right) \times 100\% = 40\%$ in Figure 8.24). For $\omega_{min} < \omega_b$, $i_{over} > 40\%$ for $F = 10$, in all designs. For lower values of F , i_{over} may be lower than 40%.

There are few essential constraints that limit the rather large constant power speed range of PM-RSM. The most important are as follows:

- emf at maximum speed
- Uncontrolled generation elimination
- Limiting the short-circuit current (torque)

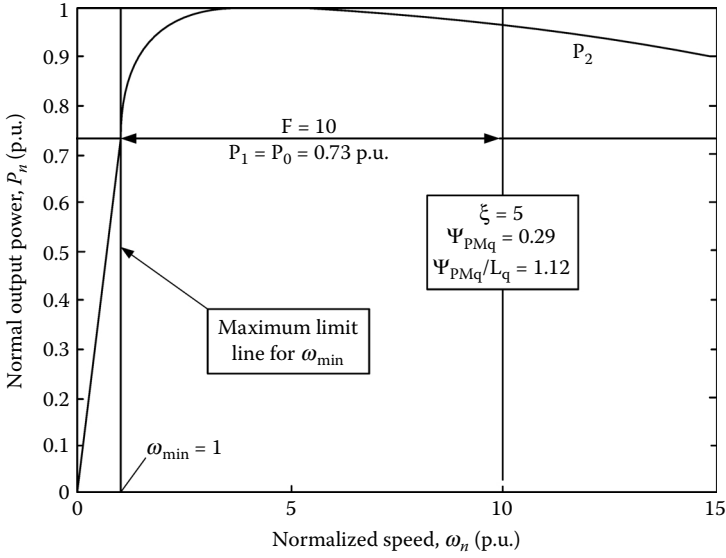


FIGURE 8.24 Design for $\omega_{\min}/\omega_b \leq 1$; $t_{\text{over}} = 0$, $i_{\text{over}} = 40\%$.

The emf at maximum speed E_{smax} is as follows:

$$e_{\text{smax}} = \frac{E_{\text{smax}}}{V_b} = \frac{\omega_{\text{max}} \Psi_{PMq}}{V_b} \quad (8.71)$$

In general, $e_{\text{smax}} \leq 150\%$ to avoid voltage overrating of the converter. This constraint may become severe for $F = \omega_{\text{max}}/\omega_b = 10$, as $\Psi_{PMq} = 0.15$ (P.U.), which is a small value. High saliency ratio $L_d/L_q > 5$ would save such a design, eventually.

The symmetrical ideal short-circuit current i_{sc3} (Equation 8.51) is as follows:

$$i_{\text{sc3}} = \frac{\Psi_{PMq}}{L_q} \quad (8.72)$$

In close to optimal designs, however, $\Psi_{PMq}/L_q = 1$, and thus, the symmetrical short-circuit current becomes equal to the base (peak torque) current I_b . Unfortunately, in asymmetrical short-circuits, the current is notably larger than $i_{\text{sc3}} \cong I_b$. The three-phase fault may occur unintentionally by gating all inverter switches or through DC bus short-circuit, when all six freewheeling diodes are conducting. Asymmetrical short-circuit through the failure of a single inverter switch in a closed position behaves as single-phase short-circuit at the machine terminals. Much higher currents than for symmetrical short-circuit occur within 6 msec in two of the three phases that also exhibit DC components. High pulsating torque values are expected, as in any synchronous machine. As a protective means from these high transients, the unfaulted phases may be opened or the machine may be driven into symmetrical short-circuit, when the overcurrents are acceptable ($i_{\text{sc3}} \cong i_b$ for $\Psi_{PMq}/L_q \cong 1$). A pertinent analysis of this problem is presented in Reference [16]. Finally, uncontrolled generator (UCG) operation may occur at high speeds when gating is removed from all the power switches. The machine acts as a generator with uncontrolled rectifier output. The problem is that due to $L_d/L_q \gg 1$, the machine continues to generate power even when the speed decreases in UCG below the speed that corresponds to $\omega \Psi_{PMq} = V_b$. It was shown [17] that, in order to avoid UCG, the following condition is to be met: $\omega_{\text{max}} < \omega_{\text{UCG}}$.

TABLE 8.2 Typical Specifications of a Starter/Alternator (ISA) for Cars

Parameter	Value
Starting torque at crankshaft (with or without a gear to ISA)	140 nm
Electric generating power: at crankshaft at 0.6 krpm	2–4 kW
Electric generating power: at crankshaft at 6 krpm	3–6 kW
Maximum starting current density	20–50 A/mm ²
Maximum generating current density	10–20 A/mm ²
Nominal bus voltage	42 V _{dc}
System efficiency at full load and 1500 rpm	75%
Minimum rotor burst speed	10–18 krpm
Maximum electromagnetic field (emf) at 6 krpm	24 V _{ph} (rms)

With

$$F = \frac{\omega_{UCG}}{\omega_b} = 2 \frac{\sqrt{\frac{L_d}{L_q} - 1}}{\Psi_{PMq}(L_d/L_q)} \quad (8.73)$$

For our previous close to optimal design $\Psi_{PMq} = 0.29$, $L_d/L_q = 5$, $F = \omega_{UCG}/\omega_b = 2.75$.

Only by adopting much smaller values of Ψ_{PMq} and a higher saliency ratio may a larger constant power speed range be produced, at the cost of high inverter oversizing. The maximum emf constraint, for the case in point, fulfilling $\Psi_{PMq} = 0.15$ (P.U.) for $F = 10$, fails to produce an L_d/L_q ratio (even a very large one) in Equation 8.73. Only for $\Psi_{PMq} = 0.1$ for $F = 10$, does $L_d/L_q = 2$ produce a solution to Equation 8.73. Unfortunately, this is so small that the machine torque density is unacceptable. It seems that elimination of UCG in large constant power speed range design ($F > 2.75$), with PM-RSM, is difficult to achieve unless the PM flux is very small and saliency is very high. But then, all the merits of PM-RSM are diminished drastically as the machine turns into an RSM.

UCG operation has to be handled by some other protective means, as the constraint $\omega_{max} < \omega_{UCG}$ for its elimination does not look practical for industrial designs. As shown in this paragraph, though the PM-RSM seems capable of producing 10:1 constant power speed range with small machine oversizing and 40% converter oversizing, it fails to comply with 150% maximum speed emf conditions and especially fails to avoid UCG at high speeds. Designing PM-RSM for a 4:1 constant power speed range and making use of a star to delta winding switching, as done for the induction machine seems to be the way to go. Machine oversizing and converter oversizing and the total system losses are about 30 to 40% less than for an induction starter alternator designed for the same values. Magnetic saturation reduces the saliency ratio L_d/L_q with all the consequences that follow [18]. Typical specifications for integrated starter alternators (ISA) for cars are shown in Table 8.2.

Offset coupling of ISA to the crankshaft may be preferred. With a 3:1 gear (belt) ratio, the PM-RSM, with an overall slightly expensive design, may be obtained with unlimited emf, as the peak current is lower in the 2:1 ratio. Consequently, the converter voltage oversizing is compensated for by the current undersizing [19]. Tripling the speed for geared PM-RSM, however, may impose more severe mechanical constraints on the design.

8.7 Power Electronics for PM-RSM for Automotive Applications

As the PM-RSM in automotive applications undergoes cyclic energy exchanges with a battery, a PWM, voltage source inverter capable of bidirectional power flow is required (Figure 8.25). The capacitor filter helps reduce the current transients in the battery during PWM converter switching sequences. The power switches have to be metal-oxide semiconductor field-effect transistors (MOSFETs) in 42 V_{dc} battery

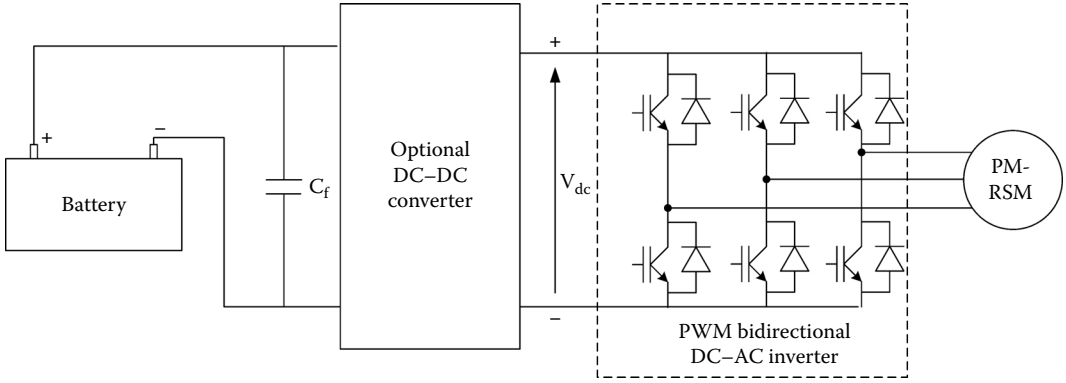


FIGURE 8.25 Typical pulse-width modulator (PWM) converter for permanent-magnet-assisted reluctance synchronous machine (PM-RSM).

systems with voltage boost via an optional DC–DC converter (Figure 8.25). Introducing the DC–DC converter for voltage boost allows for design of the PM-RSM and the PWM converter for higher voltage, low current ratings. Also, the DC voltage V_{dc} is made variable and the PWM converter may work extensively in the six-pulse mode when the commutation losses are reduced and the maximum modulation index is increased by 5% in comparison with standard PWM techniques (without overmodulation).

There are additional losses in the DC–DC converter. However, the possibility of using insulated gate bipolar transistor (IGBTs) in the higher-voltage PWM inverter leads to cost reductions in comparison with MOSFETs, but there are the additional costs of the DC–DC converter that have to be added.

The large battery voltage variations of $\pm 20\%$ or more may be absorbed by the DC–DC converter, allowing for designing the PWM inverter at least at maximum battery voltage, V_{bmax} .

The typical kVA_{ref} of the PWM three-phase inverter to deliver given active power is as follows:

$$KVA_{ref} = \frac{\sqrt{3}}{2} V_{max} I_{max} \quad (8.74)$$

where

I_{max} is the peak phase current value

V_{max} is the peak line voltage value

The inverter component size A_{inv} is defined through the $I_{dim} = I_{max}$ and V_{dim} :

$$A_{inv} = 6V_{dim} I_{dim} \quad (8.75)$$

In general, $V_{max} < V_{dim}$. When a DC–DC converter is present, a safe value of V_{max} for 600 V IGBTs would be $V'_{max} = V'_{dc} = 400 V_{dc}$ with $V'_{max} = V_{bmin} = 125 V_{dc}$ when the DC converter is missing.

For a 42 V_{dc} system, MOSFETs are to be used, and again, if a DC–DC converter is present, $V'_{max} \cong 3 V_{bmin}$. (V_{bmin} is the minimum battery voltage.) The ratio between the inverter size and delivered mechanical power in motoring is as follows [2]:

$$\frac{A_{inv}}{P_n} = \frac{6V_{dim} I_{dim}}{P_n} = 4\sqrt{3} \frac{V_{dim}}{V_{max}} \cdot \frac{KVA_{ref}}{P_n} \quad (8.76)$$

The DC–DC converter size is defined as follows:

$$A_{dc-dc} = 2V_{dim} I_{bmax} \quad (8.77)$$

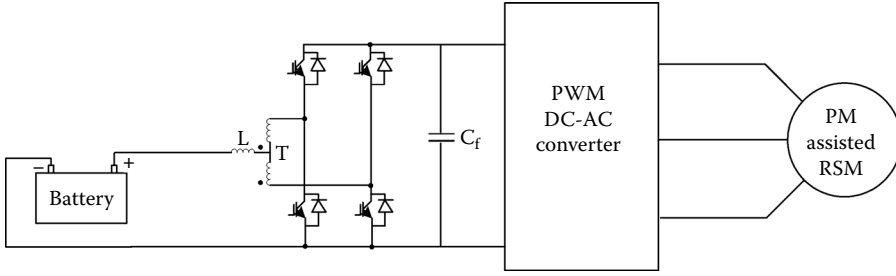


FIGURE 8.26 H-bridge direct current (DC)–DC boost bidirectional converter with transformer and inductance ($T + L$).

where $I_{b\max}$ is the maximum battery current.

Adapting a total (system) efficiency η_{tot} yields the following:

$$\frac{A_{dc-dc}}{P_n} = \frac{2}{\eta_{tot}} \frac{V_{d\min}}{V_{b\min}} \quad (8.78)$$

The ratio between the inverter plus DC–DC converter to “inverter-only” size is

$$\frac{A_{dc-dc} + A_{inv}}{A_{inv}} = \frac{V_{b\min}}{V'_{\max}} + \frac{1}{2\sqrt{3}\eta_{tot}} \cdot \frac{P_n}{KVA_{ref}} \quad (8.79)$$

The ratio P_n/kVA_{ref} varies between 0.25 and 0.6, while $(1/2\sqrt{3}\eta_{tot})$ is roughly $0.25 \div 0.35$. For $1/2\sqrt{3}\eta_{tot} = 0.34$ and $P_n/kVA_{ref} = 0.25$, the second term in Equation 8.78 is 0.2; thus, even at $V_{b\min}/V'_{\max} < 0.8$, the total size of the DC–DC PWM converter will be lower than that of the PWM inverter at lower voltage, acting alone, and so will be the costs if the additional power switch control devices costs are small.

Even if the $V'_{dc} = V_{b\max}$ (maximum battery voltage), it seems worth adding the DC–DC converter in terms of initial silicon costs. The DC–DC boost converter includes inductances for energy storage and voltage boost, but the size of the capacitive filter in front of the PWM inverter is reduced [2] and so is the current ripple in the battery. A typical H-bridge DC–DC boost converter with inductance and transformer is shown in Figure 8.26 [2].

Another key issue is the total loss in power electronics. Converter losses are made of conduction and switching losses. For an IGBT inverter leg (two switches),

$$(P_{cond})_{leg} = \left(V_I (I)_{avg} + \frac{\Delta V_R}{I_{rated}} (I)_{RMS}^2 \right) \quad (8.80)$$

$$(P_{sw})_{leg} = \beta_{sw} V_{DC}^I (I)_{avg}$$

The switching losses per lag in the PWM inverter for six-pulse mode (hexagonal, voltage 5% overmodulation) are as follows [2]:

$$P_{sw6pulse} = \frac{\sin|\varphi|}{2} P_{swPWM}; \quad |\varphi| > \frac{\pi}{6}$$

$$P_{swHpulse} = \left(\frac{2 - \sqrt{3} \cos|\varphi|}{2} \right) P_{swPWM}; \quad |\varphi| < \frac{\pi}{6} \quad (8.81)$$

The power factor angle φ influences the commutation losses, as known. Typical values of the above coefficients for IGBTs are $V_I = 0.9$ V (IGBT voltage drop), $\Delta V_R = 1.1$ V (diode voltage drop), and $\beta_{sw} = 0.042$.

It was demonstrated that total losses in the power electronics with boost DC–DC converters, for 30 kW, $V_{b\min} = 125$ V_{dc} ($V_b = 180$ V_{dc}), $V_{\max} = 400$ V [2], are smaller than those of the PWM inverter connected

directly to the battery terminals. And, again, the PM-RSM is better in terms of total losses — machine plus power electronics — than all the other solutions for uphill or flat-terrain driving. In addition, the kVA_{ref} of a PM-assisted RSM converter is smaller, and thus, the cost of power electronics is smaller.

8.8 Control of PM-RSM for EHV

Essentially, as for induction starter/alternators (Chapter 7), the control of PM-RSM in EHV's does torque control from zero speed and has to operate with a large constant power speed range. It has to accommodate the driver's motion expectations while observing the best possible energy transfer to and from the battery. And, all this in starter (motoring) generating — for regenerative braking and for battery recharge — and again in motoring the air conditioner or other auxiliaries during frequent engine idle stops.

As for the induction machine, FOC or direct torque and flux control (DTFC) are both feasible strategies. Also the control system may use motion feedback (a position sensor) or may be motion sensorless (without motion feedback).

The control system has to operate safely and nonhesitantly in torque control – for motoring – from zero speed. So, in contrast to the induction machine, the initial rotor position has to be estimated first, in motion sensorless control. Typical control schemes for FOC and DTFC with motion (position) feedback are shown in Figure 8.27 and Figure 8.28 and, respectively, in Figure 8.29 and Figure 8.30.

As explained earlier, in EHV applications, torque control prevails, and thus, even for generating, the reference torque (negative) is to be used. The operation modes are as follows:

- Starting (motoring)
- Regenerative braking (generating)
- Battery slow recharging (generation)
- Motoring during engine idle stop

The battery state characterized here by the battery voltage only (but more complex in general) has to be considered in limiting the reference torque. Also, the speed will set limits on the available torque, as the inverter output voltage is limited. The torque limitation with speed may be done many ways, but from maximum torque per current (up to base speed) to maximum torque/stator flux at maximum speed, the current d - q angle may be varied as planned off-line (Figure 8.28). Only the motoring is treated in Figure 8.28 for i_d^* , i_q^* calculators.

Expressions similar to those in Figure 8.28 may be derived for generating, but for (eventually) more flux in the machine (for given torque). The voltage decoupling is required to secure better control at

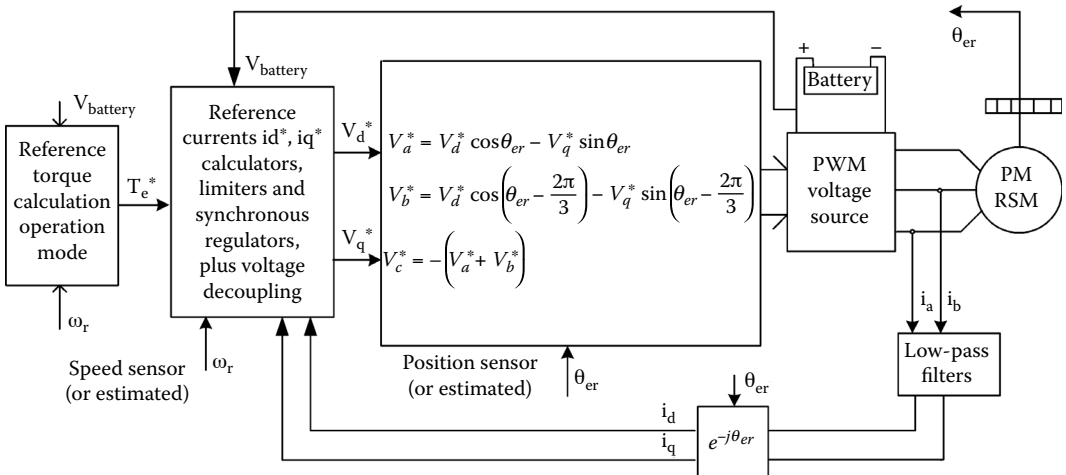


FIGURE 8.27 Indirect current voltage vector control with synchronous current regulators.

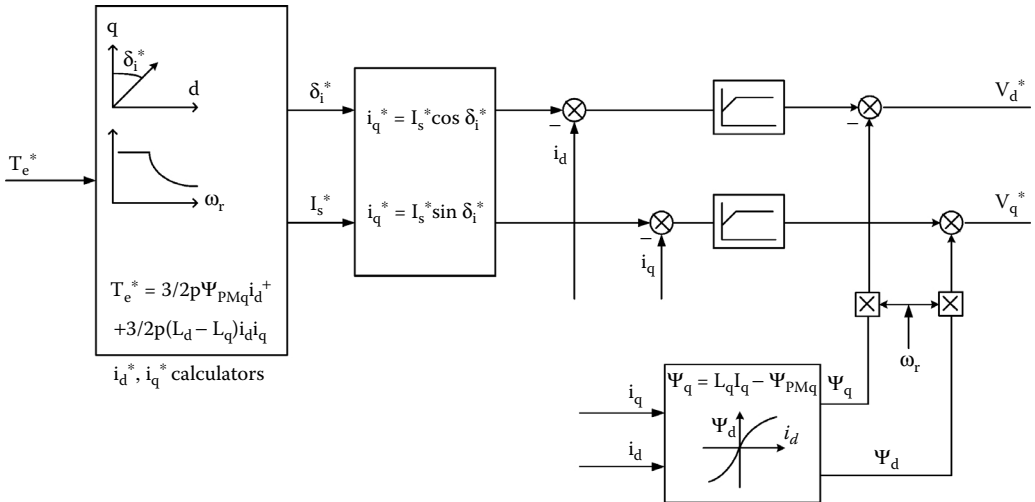


FIGURE 8.28 Reference currents i_d^* , i_q^* calculator and synchronous regulators with voltage decoupling.

high speeds, where the synchronous current regulators have less room for adequate control due to the influence of motion-induced voltage.

The presence of an absolute position sensor encoder provides for safe start from any initial position and allows for speed calculation down to less than 1 rpm through adequate digital filtering. The precalculation of optimum current angle δ_i^* as a function of speed (eventually even of reference torque) by gradually advancing it with speed, based on maximum torque/flux condition, simplifies the control notably, as the current amplitude i_s^* and d - q angle δ_i^* are found unequivocally once the reference torque and the actual speed are known.

A DTFC system for PM-RSM with position feedback is shown in Figure 8.29. Provided the stator flux and torque estimates are fast, robust, and precise and work from zero to maximum speed, DTFC provides for more robust control with about the same computation effort. The absence of reference online calculators of i_d^* , i_q^* and of the two vector rotators in DTFC is somewhat compensated for by the occurrence of the stator flux and torque estimator.

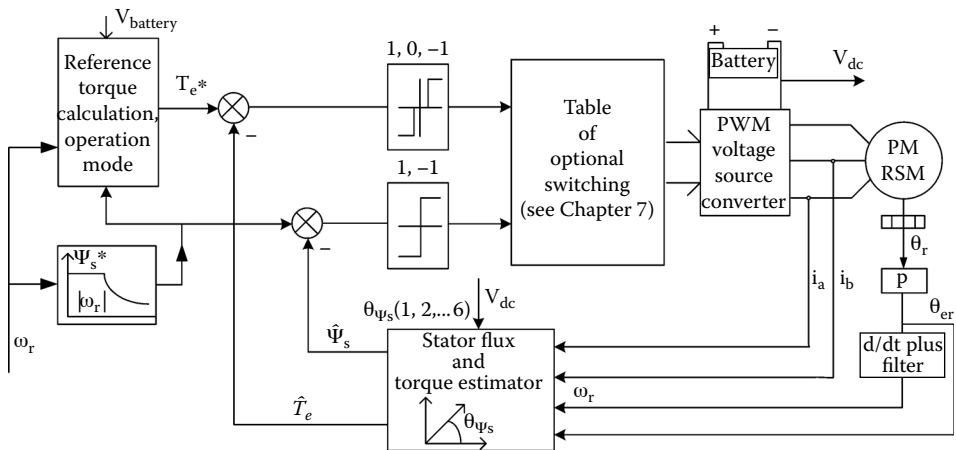


FIGURE 8.29 Direct torque and flux control (DTFC) of permanent-magnet-assisted reluctance synchronous machine (PM-RSM) with position feedback.

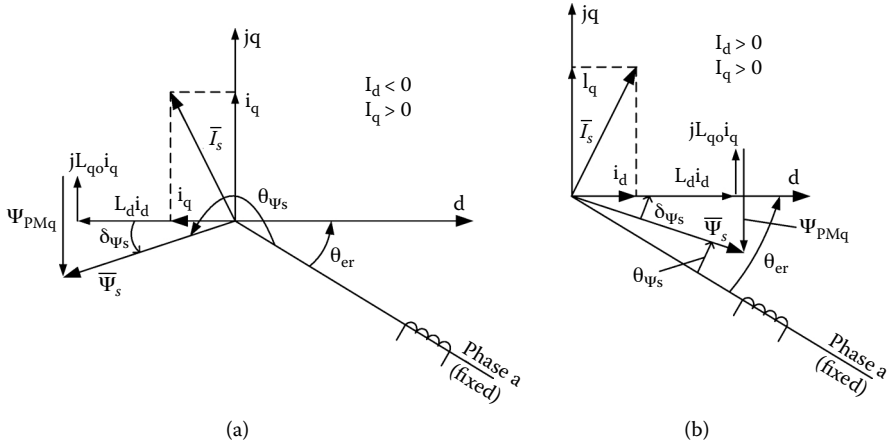


FIGURE 8.31 Rotor position θ_{er} in relation to δ_{ψ_s} and θ_{ψ_s} : (a) generating and (b) motoring.

$$\bar{\theta}_{\psi_s} = \sin^{-1} \left(\frac{\Psi_s^s}{|\bar{\Psi}_s|} \right); \quad \Psi_d = \Psi_s \cos \delta_{\psi_s} \quad \Psi_q = \Psi_s \sin \delta_{\psi_s} \quad (8.84)$$

$$T_e = \frac{3}{2} p \left(\Psi_{PMq} + (L_d(\Psi_d) - L_{q0}) \frac{\Psi_s \sin \delta_{\psi_s}}{L_{q0}} \right) \frac{\Psi_s \cos \delta_{\psi_s}}{L_d(\Psi_d)} \quad (8.85)$$

The stator flux and torque observer in Figure 8.30 may be augmented with Equation 8.83 through Equation 8.85 to include the rotor position θ_{er} calculator. Equation 8.85 may be simplified, as δ_{ψ_s} is, in general, a small angle (below 30° in most cases), and $\cos \delta_{\psi_s} \approx 1$:

$$|\hat{\delta}_{\psi_s}| \approx \left(\frac{-2|\hat{T}_e| L_d(\hat{\Psi}_s)}{3p \hat{\Psi}_s^2} + \frac{\Psi_{PMq}}{\Psi_s} \frac{L_d(\Psi_s)}{L_q} \right) \frac{1}{\frac{L_d(\Psi_s)}{L_{q0}} - 1} \quad (8.86)$$

This way, the online computation effort is reduced drastically. Such a position estimator θ_{er} (calculator) is acceptable for DTFC schemes, as θ_{er} appears only in the flux observer. This is not so for FOC schemes, where θ_{er} is crucial for field orientation.

Still, the speed has to be estimated. Again, as precise speed control is not needed, the estimated speed is used for adjusting the reference flux Ψ_s^* and reference torque T_e^* . At low speeds, this information is not important, as Ψ_s^* generally remains constant. To a first approximation, then, the steady-state speed estimation may be adopted. That is, the stator flux vector speed is used instead of $\hat{\omega}_r$:

$$\hat{\omega}_{\psi_s} \approx \hat{\omega}_r = \frac{\Psi_{s\beta}(k)\Psi_{s\alpha}(k-1) - \Psi_{s\alpha}(k)\Psi_{s\beta}(k-1)}{T_{sample}(\Psi_{s\alpha}^2(k) + \Psi_{s\beta}^2(k))} \quad (8.87)$$

After low-pass filtration, such a signal should be enough for torque and flux references. So, for sensorless DTFC of PM-RSM, with torque (not speed) control at zero speed, the flux and torque estimators need rotor position and speed calculators that are fast but not so accurate.

The question remains if such a fundamental excitation (no signal injection) state observer can provide nonhesitant start and fast torque control at zero speed. At least for the induction machines, it did. Many other state observers for FOC or DTFC of IPM synchronous machines without signal injection were proposed. Luenberger, MRAS, and Kalman filter observers all use the stator voltage vector equation

(voltage model) and the current model, and the flux current relationship in d - q rotor coordinates. Sliding mode such observers were proven to yield good stator flux rotor position and speed estimation down to very low speeds (a few rpm) [20]. However, robust fast torque sensorless control of PM-RSM at zero speed, without signal injection, is still due to be proven by tests. And, this is so due to the need to estimate first, at zero speed, the initial rotor position. This is how the signal injection state observers for PM-RSM come into play.

8.10 Signal Injection Rotor Position Observers

Signal injection state observers are based on machine magnetic saliency tracking. Back-emf methods do not use signal injection but cannot work at very low speeds, or at zero speed. These methods are used to detect rotor position, and then, from the latter, the rotor speed is estimated through adequate filtering. Saliency tracking by signal injection works well from zero speed, but it limits available DC link voltage at high speeds. A combination of the saliency tracking — to 10% maximum speed — and back-emf tracking above 10% of maximum speed was recently proven experimentally as being very good from 0 to 5000 rpm in 70 kW IPM synchronous machine control [21] with $\pm 100\%$ torque and 150 msec ramp control at 100 rpm and at 5000 rpm.

Saliency tracking methods are used to estimate absolute rotor position. A voltage (current) signal at a carrier frequency ω_c above 500 Hz is injected in rotor d - q coordinates in the FOC. The response in current is used to track the spatial saliency and thus the rotor position. The injected signals (in voltage) are of the following form:

$$\bar{V}_{dq}^c = V_c \cos \omega_c t - j V_c \sin \omega_c t \quad (8.88)$$

The machine response in current, separated through adequate filtering, is as follows:

$$\bar{I}_{dq}^c = I_p \sin \omega_i t - I_n \cos(h\theta_{er} - \omega_i t) - j(I_p \cos \omega_i t - I_n \sin(h\theta_{er} - \omega_i t)) \quad (8.89)$$

I_p and I_n are positive and negative sequence carrier currents, while h is the order of the saliency spatial harmonic; θ_{er} is the rotor position. Only the negative sequence current contains the rotor position information. Also, in general, only one spatial harmonic (h) is considered. The negative sequence current is processed through heterodyning to produce a signal that is proportional to the rotor position error. This $\Delta\theta_{er}$ (error) signal is then used as input to a Luenberger observer that produces parameter-insensitive zero lag position and speed estimates [22–24].

For the back-emf methods, used for high speeds, the method in Section 8.9 is typical. The basic structure of such a hybrid sensorless FOC, with saliency plus back-emf position and speed tracking, is shown in Figure 8.32 [21]. The switching from low-speed position estimation to high-speed position estimation is done when the two have the same output, around 10 to 15% of maximum speed, to leave room for full voltage utilization by the voltage source inverter. Position estimation results at 100 rpm and at 5000 rpm, for a 70 kW IPM-SM, obtained with the above combined control method, is shown in Figure 8.33 and Figure 8.34 [21].

Torque ranging of $\pm 100\%$ is experienced with apparently flowless position estimation. Still, for safe torque control at zero speed, the initial rotor position is required. This is not directly possible with the above method.

8.11 Initial and Low Speed Rotor Position Tracking

A simple way to detect initial rotor position is to use three proximity Hall sensors placed in the axes of the three phases, to detect the PM flux position. The resolution of such a coarse position tracker is only $180/3 = 60^\circ$ (electrical). For DTFC, such an error may be acceptable, as the method still indicates correctly

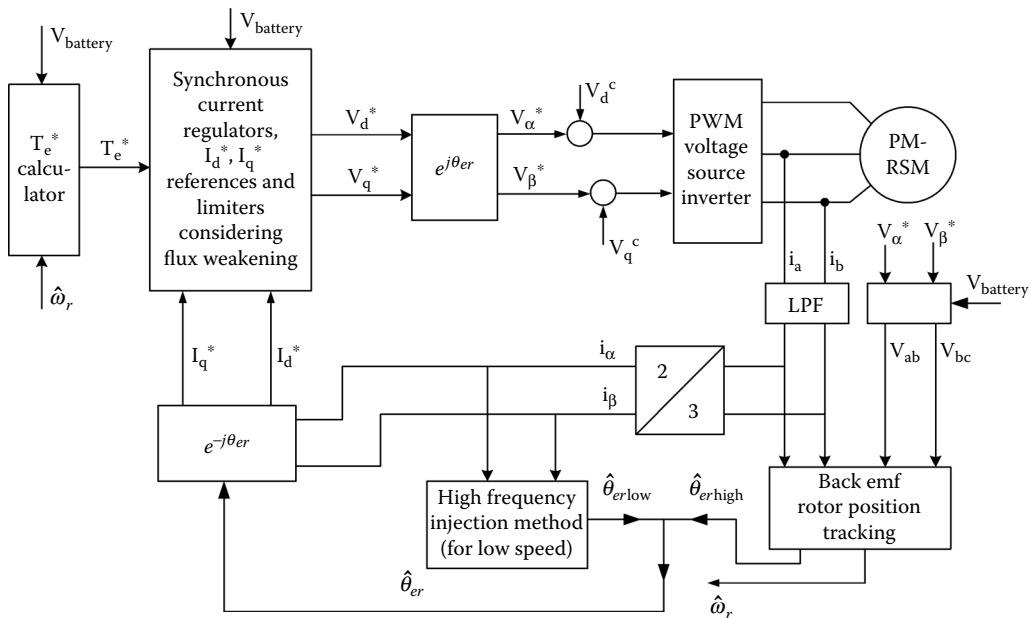


FIGURE 8.32 Sensorless flux-oriented control (FOC) of permanent-magnet-assisted reluctance synchronous machine (PM-RSM) with signal injection plus back electromagnetic force (emf) rotor position estimation.

the first voltage vector in the table of optimal switchings to be triggered in the inverter. For FOC, better precision is required to secure safe smooth torque for starting. The additional wiring for the Hall sensors may lead to failure. Pulse signal [25] or sinusoidal carrier signal [23–29] methods were proposed in the last decade for initial position tracking in IPM-SMs.

The slight magnetic saturation due to the PM “North Pole” presence is used to find the PM polarity and, thus, settle for the actual rotor initial position (Figure 8.35). Consequently, magnetic saturation

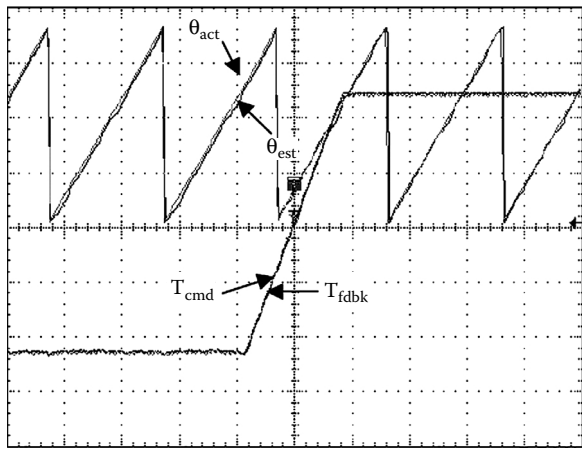


FIGURE 8.33 Dynamic performance of injection at 100 rpm. (Adapted from M. Patel, T. O’Meara, J. Nagashinia, and R.D. Lorenz, Encoderless IPM drive system for EH-HEV propulsion applications, Record of EPE-2001, Graz, Austria, 2001.)

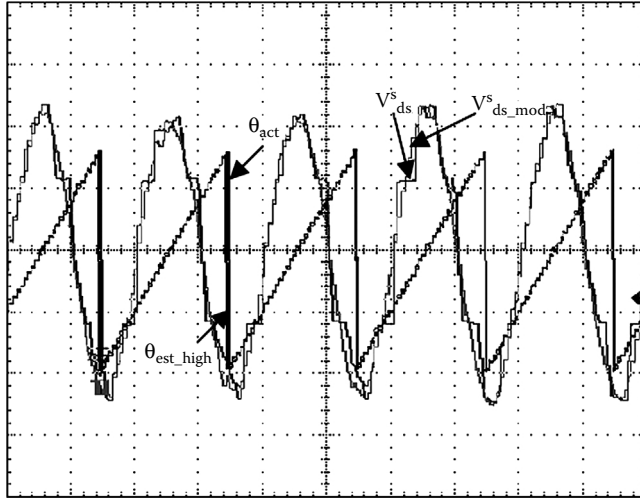


FIGURE 8.34 Comparison of actual and estimated rotor position at 5000 rpm in six-step mode under full load condition using BEMF method. (Adapted from M. Patel, T. O'Meara, J. Nagashinia, and R.D. Lorenz, Encoderless IPM drive system for EH-HEV propulsion applications, Record of EPE-2001, Graz, Austria, 2001.)

has to be considered in the PM-RSM model. The voltage model vector equation of PM-RSM is as follows:

$$\bar{V}_s = R_s \bar{i}_s + \frac{d\bar{\Psi}_r}{dt} + j\omega_r \bar{\Psi}_s \quad (8.90)$$

$$\bar{\Psi}_s = \Psi_d + j\Psi_q \quad (8.91)$$

Magnetic saturation is considered separately for the two axes:

$$i_d + ji_q = R_{d0}\lambda_d + R_{d1}\lambda_d^2 + R_{d2}\lambda_d^3 + j[R_{q0}(\lambda_q - \lambda_{PMq}) + R_{q1}(\lambda_q - \lambda_{PMq})^2] \quad (8.92)$$

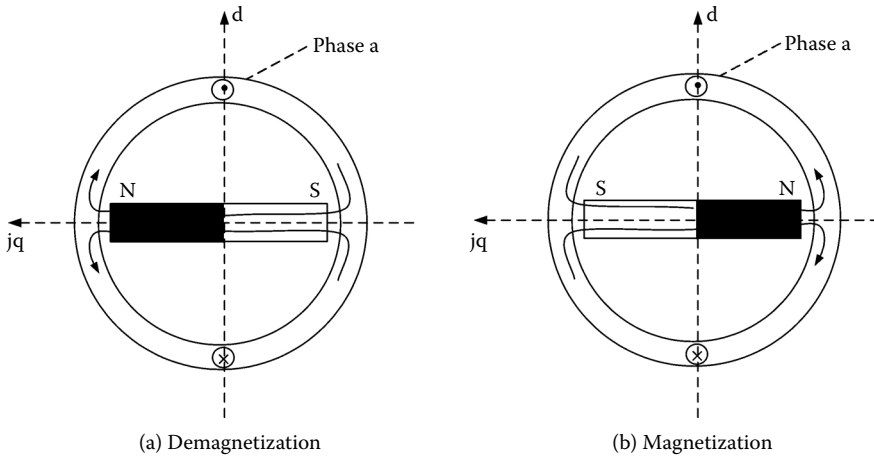


FIGURE 8.35 Magnetic saturation in the q axis of a permanent-magnet-assisted reluctance synchronous machine (PM-RSM).

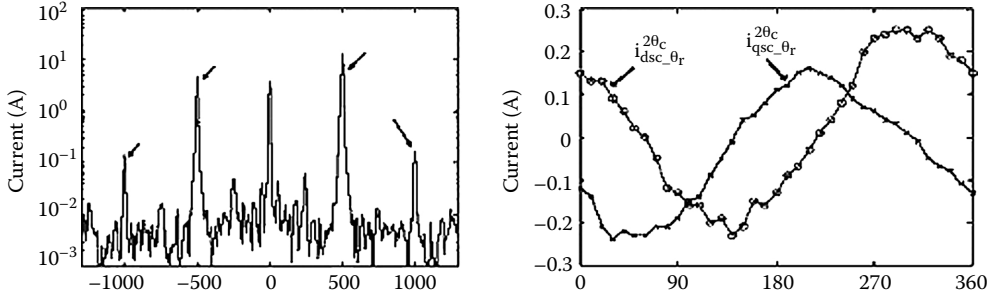


FIGURE 8.36 Measured carrier current spectrum.

Only two terms were retained in axis q , as the magnetic saturation here is less important, as demonstrated earlier in this chapter. Along axis d , magnetic saturation is heavy for the large starting torque. The magnetic saturation functions are mere qualitative, used to discriminate the machine current response waveform when a sinusoidal voltage signal injection is applied. A sinusoidal voltage injection method is illustrated in what follows [30].

The injected signal $V_{\alpha\beta}^c$ in stator coordinates is as follows:

$$V_{\alpha\beta}^c = V_c e^{j\omega_c t} \quad (8.93)$$

In rotor coordinates,

$$V_{dq}^c = V_{\alpha\beta}^c e^{-j\theta_{er}} \quad (8.94)$$

Neglecting the resistive voltage drop,

$$\bar{\Psi}_{dq}^c = \int \bar{V}_{dq}^c dt \quad (8.95)$$

Then, making use of Equation 8.92, the injected currents in stator coordinates are as follows:

$$\bar{I}_{\alpha\beta}^c = \bar{I}_{\phi 1} e^{j\left(\omega_c t - \frac{\pi}{2}\right)} + \bar{I}_{cn1} e^{j(-\omega_c t + 2\theta_{er} + \phi_{1n})} + \bar{I}_{\phi 2} e^{j(2\omega_c t - \theta_{er} + \phi_{2p})} + \bar{I}_{cn2} e^{j(-2\omega_c t + 3\theta_{er} + \phi_{2n})} \quad (8.96)$$

The first and second time harmonic components are illustrated in Equation 8.96. Only the last three terms contain rotor position information at the second, first, and, respectively, third spatial harmonic. The measured carrier current spectrum at standstill in stator coordinates (Equation 8.96) is shown in Figure 8.36 [30]. The negative sequence first time harmonic (second-space harmonic) current — the second term in Equation 8.96 — contains the largest signal and seems to be most appropriate for initial position detection.

In Reference [30], the I_{cn1} and I_{cn2} terms in Equation 8.96 are used to estimate the position error $\theta_{er} - \hat{\theta}_{er}$, after frame transformations from stationary to ω_c and $2\omega_c$ reference frame speeds, and after taking their dot product:

$$\bar{I}_{dq2\theta_{er}}^{-\theta_c} = I_{cn1} e^{j(2\theta_{er} + \phi_{1n})}; \quad \bar{I}_{dq3\theta_{er}}^{-2\theta_c} = I_{cn2} e^{j(3\theta_{er} + \phi_{2n})} \quad (8.97)$$

$$\bar{I}_{dq\theta_{er}}^s = I_{dq2\theta_{er}}^{*-\theta_c} \cdot I_{dq3\theta_{er}}^{-2\theta_c} = I_{cn1} I_{cn2} 2e^{j(\theta_{er} + \phi_{1n} - \phi_{2n})} \quad (8.98)$$

As the above method can be applied as it is at small speeds also, it may be used successfully in combination with the back-emf method for FOC (Figure 8.32) to cover the entire speed range from standstill). The same procedure may be used in DTFC to detect θ_{er} for the use in the flux observer only (Figure 8.30).

8.12 Summary

- A high magnetic saliency ($L_d/L_q > 3$) RSM with multilayer interior PMs per rotor poles in axis q is called PM-assisted RSM or PM-RSM. It may also be called a special interior PM synchronous machine.
- The reluctance torque is predominant in comparison with the PM interaction torque, for peak value.
- Even low remnant flux density PMs ($B_r < 0.8$ T) and, thus, lower-cost PMs may be used.
- In automotive applications, typical specifications ask for high start torque up to base speed ω_b , and then a wide constant power speed range $F = \omega_{max}/\omega_b > 4/1$ is required.
- The unique combination of reluctance and PM interaction torque makes the PM-RSM particularly suitable for a wide constant power speed range both in terms of system costs and losses.
- Typical rotor topologies contain two to four (five) flux barriers per half pole, which contains parallelepiped-shaped moderate (low)-cost PMs. The rotor core is made of conventional laminations. The saliency ratio $L_{dm}/L_{qm} \approx 3$ to 10, but it greatly varies with magnetic saturation level.
- Each flux barrier ends toward airgap with a flux bridge. The bridge thickness varies between 0.5 to 1.5 mm. The smaller the bridge, the better, in terms of rotor saliency L_{dm}/L_{qm} , but, alternatively, the flux bridge thickness and length are limited by the mechanical stress allowed before rotor deformation at maximum speed and torque occurred.
- The total saliency is $L_d/L_q = 3$ to 7 for PM-RSM below 100 kW (peak torque below 400 Nm) when two to four flux barriers per half-pole are used. The stator windings leakage inductance reduces L_d/L_q to values smaller than L_{dm}/L_{qm} .
- FEM direct geometrical optimization of the rotor pole structure may lead to a uniform distribution of PM flux density in the airgap, that is, to a rather sinusoidal emf for a distributed stator winding with $q = 2,3$ slots per pole and phase.
- Also, a prime number of flux bridges per pole of five or seven, leads to low cogging torque and pulsating torque with sinusoidal currents, even without stator (rotor) skewing.
- Even for heavy saturation levels, the airgap resultant flux linkage per phase is sinusoidal if magnetic saturation is uniform in the stator and rotor teeth and yokes.
- FEM analysis brings out the fact that magnetic saturation moves the maximum torque for given (high) current toward smaller i_d and larger i_q .
- The sinusoidal airgap flux per phase under load allows for full usage of d - q models of PM-RSM for parameter, steady-state, transient, and control performance estimation.
- If FEM analysis shows that magnetic saturation in axis q (where the PMs are located) is almost negligible, while it is important in axis d , especially for high torque operation; cross-coupling magnetic saturation may, thus, be neglected in many practical cases. Saturation in axis d is accounted for by a nonlinear $i_d(\Psi_d)$ function.
- Core losses in the PM-RSM occur both in the stator and the rotor. Only space harmonics in the airgap total flux distribution produce rotor core losses under steady state. Stator current time harmonics produce core losses both in the stator and the rotor. Both space and time harmonics of the airgap field produce eddy currents in the PMs on the rotor, especially at high speeds.
- From FEM static field analysis, the core losses may be calculated. An equivalent, slightly frequency-dependent, core resistance R_C is defined for use in the d - q model.
- The d - q model flux linkages $\Psi_d = L_d(i_d)i_d$, $\Psi_q = L_q i_q - \Psi_{PMq}$ lead to a resultant torque that is proportional to i_d . When torque changes sign (from motoring to generating), it is i_d rather than i_q that changes sign.

- The PMs in axis q produce a flux that opposes $L_q i_q$; that is, a demagnetizing armature reaction occurs. The current i_q should not change polarity.
- The PMs in axis q contribute to additional torque for less Ψ_q flux and, thus, to high power factor in the machine.
- Generator no-load with nonzero core losses makes the d - q output current $i_d = i_q = 0$ but the interior currents $i_{0d}, i_{0q} \neq 0$.
- With zero current losses ($R_C = \infty$), all d - q currents in the d - q model at no load are zero, and $V_{d00} = \omega_r \Psi_{PMq}$ and $V_{q00} = 0$.
- The symmetrical steady-state short-circuit current when $R_S \neq 0$, $R_C \neq \infty$ has two components i_{dsc3} and i_{qsc3} and a corresponding torque. With zero losses, the ideal symmetrical short-circuit current $i_{sc3} = i_{qs0} = \Psi_{PMq}/L_q$.
- In most wide constant power speed range designs, $\Psi_{PMq} = L_q I_{rated}$ and, thus, symmetrical short-circuit is acceptable.
- Unsymmetrical short-circuit leads to larger than i_{sc} currents with DC components that have to be treated, eventually, through quick, intentional, triggering of all power switches in the inverter to degenerate in symmetrical short-circuit. This way, the inverter is protected.
- The design of PM-RSM for a wide speed range has to compromise machine (torque) and inverter (current) oversizing t_{over}, i_{over} . After smart normalization, only Ψ_{PMq} and L_d/L_q in P.U. remain as independent variables. In essence, design close to $\Psi_{PMq}/L_q \approx 1$ in (P.U.) leads to wide constant power speed range.
- With respect to base speed ω_b (full torque at full voltage for maximum torque per current), the minimum speed ω_{min} may be chosen either above or under it. For given constant power speed range F , choosing a $\omega_{min}/\omega_b > 1$ smaller machine and some inverter oversizing are required, but the emf at maximum speed limitation at 150% rules out such a situation with designs for $\Psi_{PMq}/L_q \approx 1$ P.U. Also, the full voltage at $\omega_b < \omega_{min}$ may lead to current regulation saturation. For $\omega_{min}/\omega_b < 1$, full machine utilization is reached, but inverter current oversizing is larger.
- Uncontrolled generation (UCG) occurs even when the speed decreases to ω_{UCG} , below that for which $\omega_r \Psi_{PMq} < V_b$ and the PWM inverter power switches are turned off and diodes act alone. To avoid UCG, the maximum speed ω_{max} should be lower than ω_{UCG} :

$$\omega_{max} < \omega_{UCG} = \frac{2\sqrt{\frac{L_d}{L_q} - 1}}{\Psi_{PMq} \left(\frac{L_d}{L_q} \right)} * \omega_b; \quad \Psi_{PMq} \text{ in P.U.}$$

- It turns out that, for $F > 4$, only for very small Ψ_{PMq} values and low L_d/L_q ratios, UCG may be eliminated. Essentially, to avoid an impractical design, UCG has to be eliminated through control (protection) means.
- For EHV's, the PM-RSM starter/alternator is connected to the battery through a PWM inverter. As the battery voltage varies with $\pm 25\%$, it turns out that a voltage booster, through a DC-DC converter, may be beneficial. Even if the voltage booster acts from V_{bmin} to V_{bmax} of battery, the total ratings and silicon costs of a DC-DC converter plus PWM inverter are smaller than those for the inverter alone at V_{bmin} . Moreover, the total system losses in the machine plus converters are smaller with a DC voltage booster.
- In EHV's, torque, rather than speed, control is required both for motoring and generating.
- FOC and DTFC strategies are feasible.
- The control may be performed with or without a position encoder.
- The control has to produce first reference d - q currents i_d^*, i_q^* vs. torque and speed functions for motoring and generating in FOC, or reference flux vs. speed and torque for DTFC.
- Varying the d - q current angle with speed from maximum torque per current criterion, below base speed, to maximum torque per flux criterion, at maximum speed, seems a good way for FOC.

- For DTFC, a flux and torque observer is required.
- The position feedback is used two times for vector rotation in FOC, while only in the flux observer for DTFC. For motion-sensorless control, position observers, capable of working from zero speed, are required.
- Initial position estimation is required with FOC, for safe starting.
- Signal injection rotor position observers were proven capable of safe zero and low speed good performance. In combination with back-emf rotor motion observers, the whole speed range is covered for FOC [31, 32].
- For DTFC, as only torque control at zero speed is required, even a rotor position and flux observer without signal injection may be adequate for the whole speed range, with initial position estimation.

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